# WIRELESS ENGINEER

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MARCH 1950

VOL. XXVII TWO SHILLINGS AND SIXPENCE No. 318

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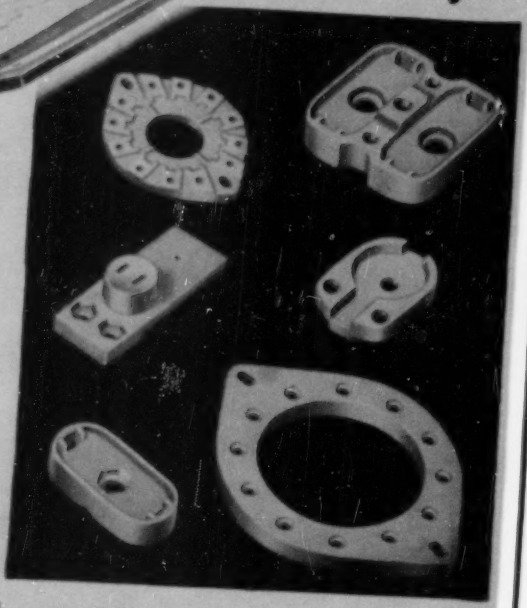

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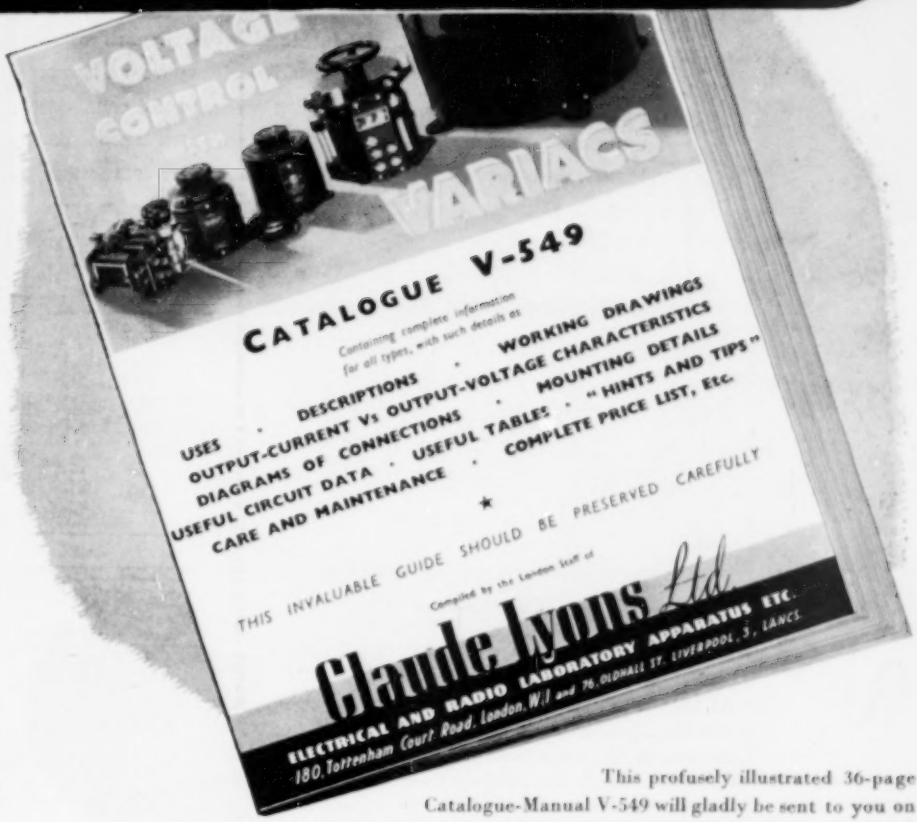
VOLTAGE
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Containing complete information for all types, with such details as
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This profusely illustrated 36-page Catalogue-Manual V-549 will gladly be sent to you on request. It gives complete data on our "Variac" Voltage Regulating Transformers together with dimensioned drawings of all types, and much useful information including many circuits and special applications and tables, etc. There is no charge for this invaluable work of reference—all applications should be addressed to :—

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IN education and in industry, these units find a host of applications There is a physical reality about them which appeals to the elementary student and which is necessarily missing from the more expensive boxes where switches are hidden mysteries and connexions must be taken on trust. This unit form of construction is economical and very convenient, and the moulded containers are sensibly engineered, so that fixing to a panel or to each other is quite simple.

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PRECISION MUIRHEAD ELECTRICAL INSTRUMENTS

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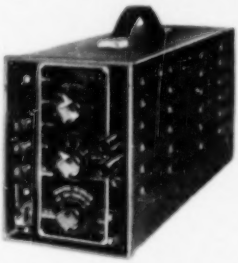
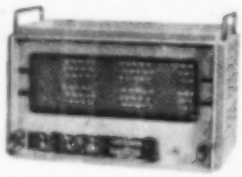
# Vortexion

## TYPE C.P.20A AMPLIFIER



For AC Mains and 12 volt working giving 15 watts output, has switch change-over from AC to DC and "Stand-by" positons. Consumes only 5½ amperes from 12 volt battery. Fitted with mu-metal shielded microphone transformer for 15 ohm microphone, provision for crystal or moving iron pick-up with tone control for bass and top. Outputs for 7.5 and 15 ohms. Complete in steel case with valves. . . . . PRICE **£28.**



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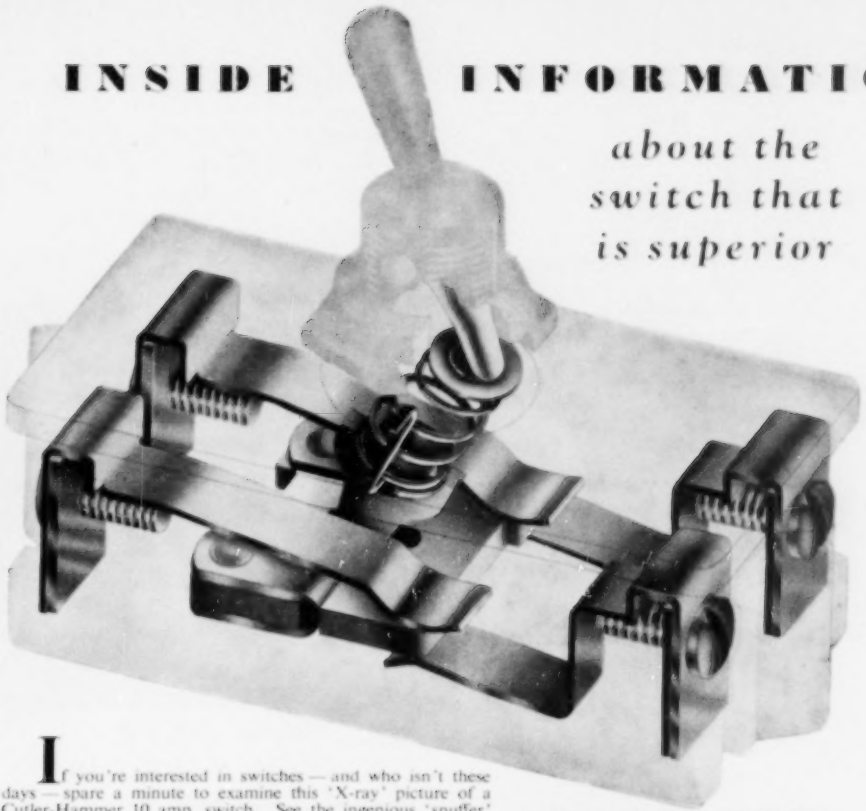
# INSIDE INFORMATION

*about the switch that is superior*



If you're interested in switches — and who isn't these days — spare a minute to examine this 'X-ray' picture of a Cutler-Hammer 10 amp. switch. See the ingenious 'snuffer' action — an original C-H feature — which eliminates pitting and arcing when breaking heavy loads. Note also that the silver alloy moving contact is self-aligning and self-cleaning. And finally observe the substantial area of the spring-leaf fixed contacts to ensure cool working.

Now if you're looking for a robust switch for a real job of work, here's the fellow to do it. May we submit samples and details of the wide range of operating mechanisms available?

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C-H

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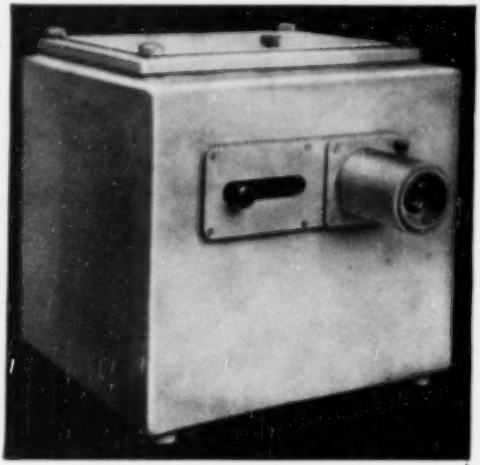

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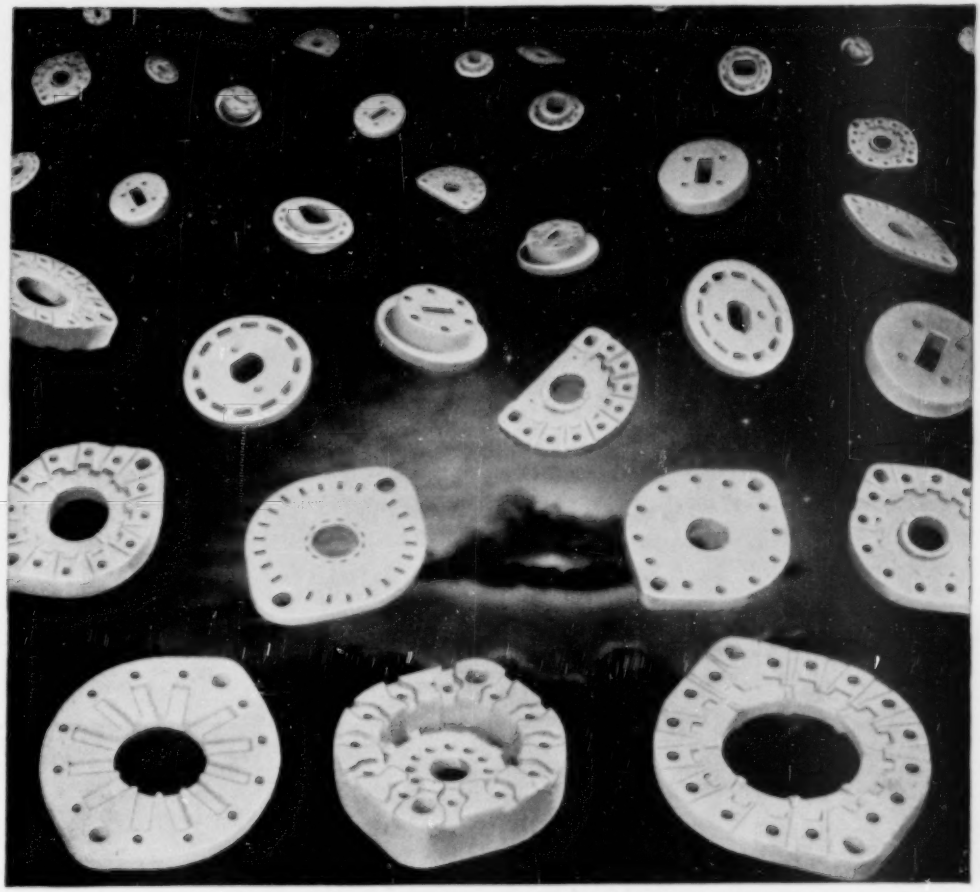

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## POLARIZED RELAYS

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### Chief Characteristics*

**High operational speeds • Freedom from contact rebound and positional error • Good contact pressure • Accuracy of signal repetition • Sensitivity of the order of 0.12 Mw. D.C. • Ease of contact adjustment • Contact gap a function of input power.**

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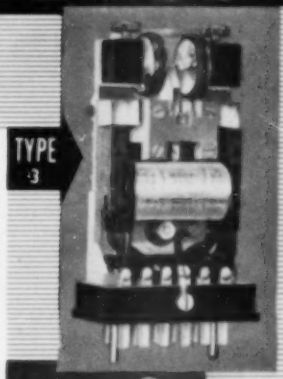
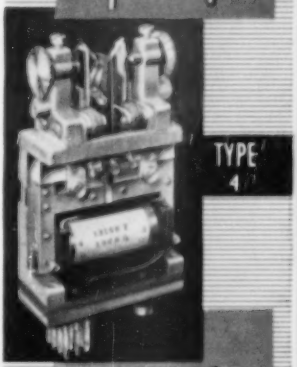
*Three Main Types are available.*




### TYPE 3

**has maximum speed : maximum contact pressure and sensitivity : minimum transit time.**
***Dimensions*** **: 120mm x 61mm x 33mm Weight 625 gm.**




### TYPE 4

**has longer contact travel than the Type 3, with medium speed and sensitivity. Also available in double-pole, double throw version with self-synchronising contacts.**
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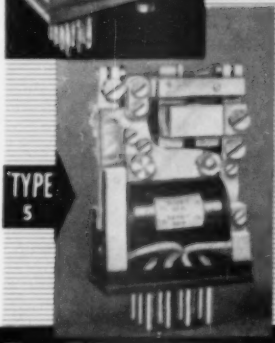


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**miniature relay with outstanding performance for size. Rugged design of exceptional thermal stability.**
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*Note : All dimensions quoted exclude connection terminals.*

• ***Complete details and performance data will be willingly supplied on request to the Manufacturers.***

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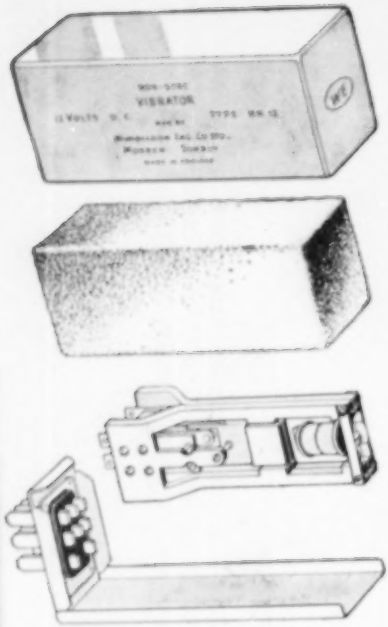

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# FREQUENCY SHIFT EQUIPMENT

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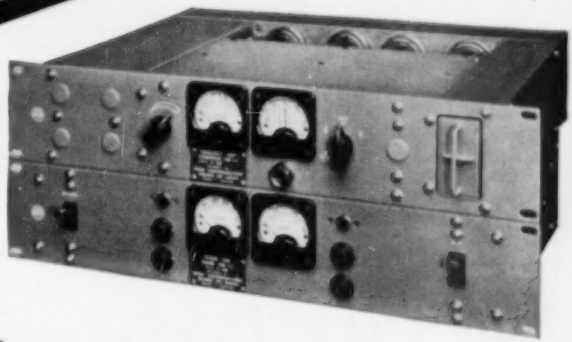

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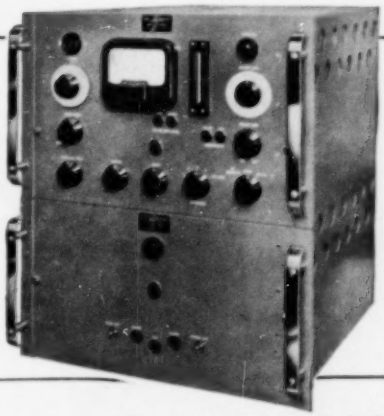

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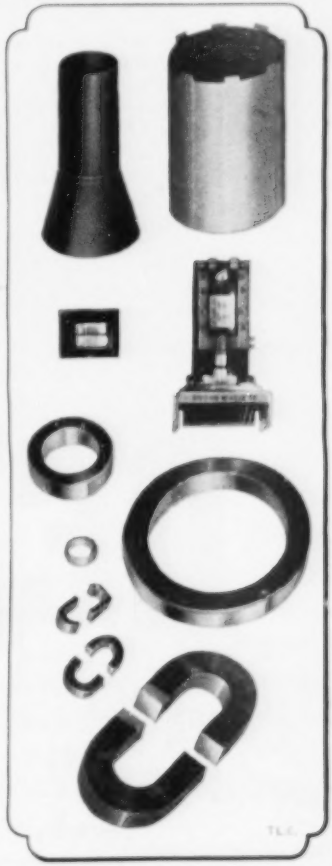

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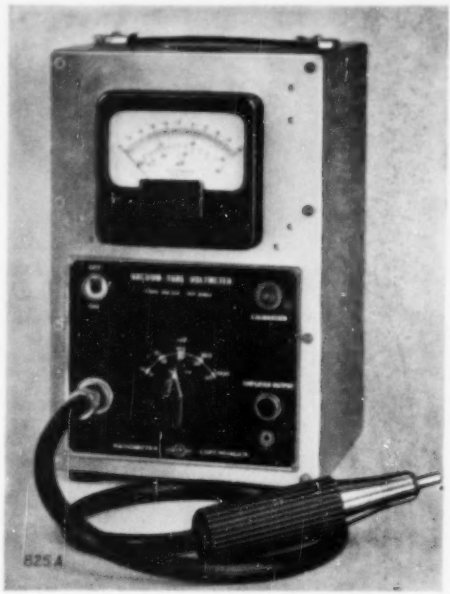


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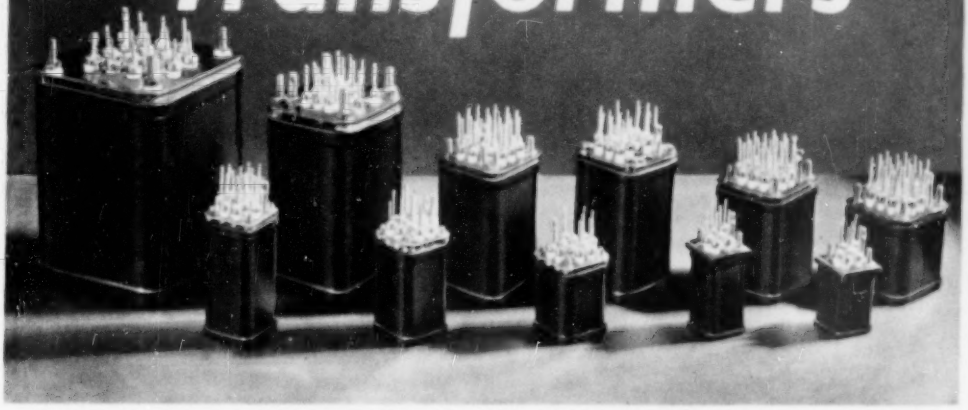
# Ferranti

MINIATURISED HERMETICALLY SEALED

# Transformers



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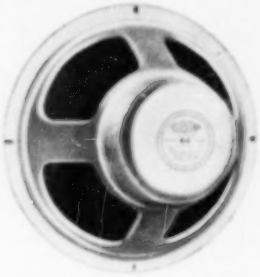

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- Overall Diameter 12 3/16"
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- Flux Density (Gauss) 10,000
- Total Gap Flux (Maxwells) 60,000
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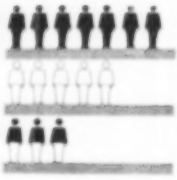
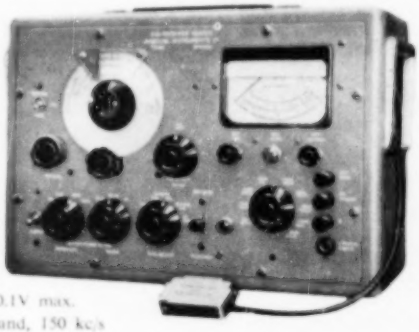


## VITAL STATISTICS



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- Crystal standardised c.w., f.m. and a.m. signal generator with audio frequency power meter.
- Carrier range 21-168 Mc/s in three bands.
- Output 1½V to 31.6 mV, continuously variable ; high level 0.1V max.
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- Amplitude modulation to nominal depth of 30% over the complete range.
- Internal modulation frequency, 1000 c/s.

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**F.M. RECEIVER TESTER** Type TF 913

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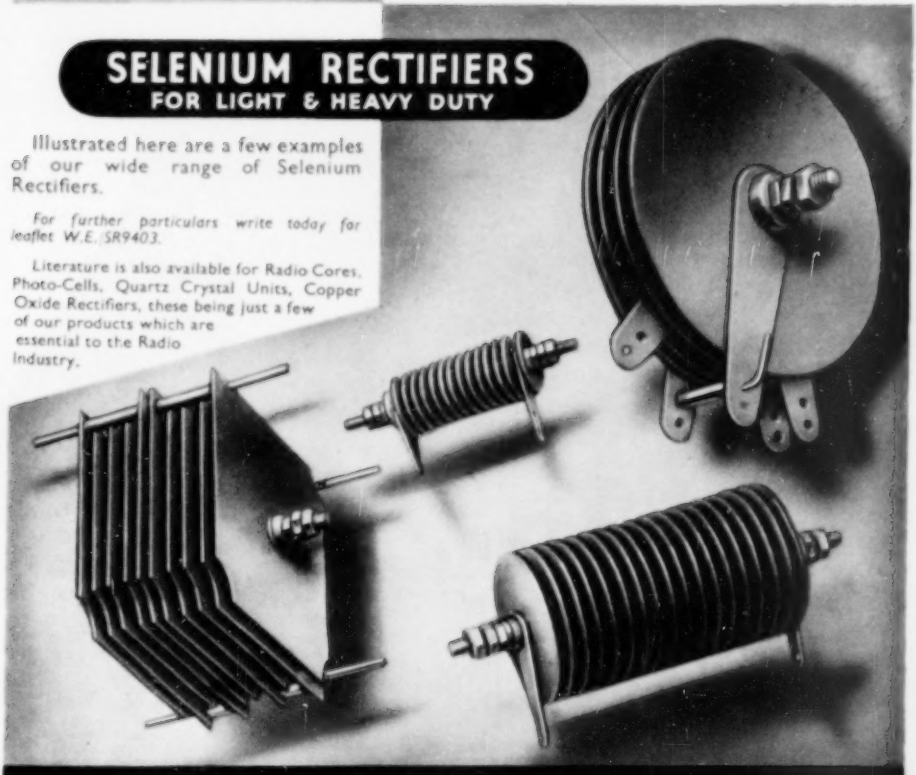

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**The Journal of Radio Research and Progress**



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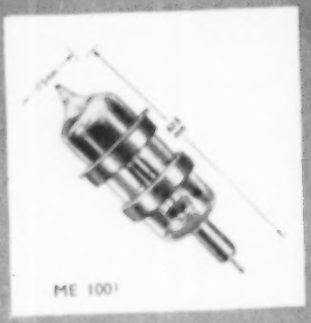


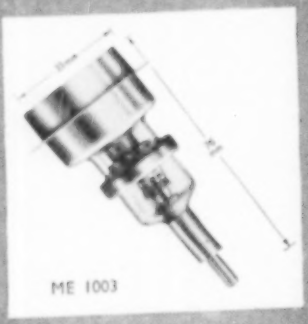


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| Max. Anode Current (mA) | 50 | 200 |
| Max. Peak Anode Current (mA) | 150 | 500 |
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MVT 64

# WIRELESS ENGINEER

Vol. XXVII MARCH 1950 No. 318


## Selectivity in Television

THE expansion of the British television service is creating some selectivity problems for designers. In view of its reputation for selectivity, therefore, it is not surprising that many are considering the possibility of the superheterodyne being more suitable than the t.r.f. set so widely used hitherto.

Most people, if asked which of the two types of receiver is the more selective, would unhesitatingly say the superheterodyne. A few would probably say that there is no difference between them. We shall show that the answer depends on circumstances and that the important factor is the relation of the product of bandwidth and circuit $Q$ to signal frequency.

We are primarily interested in the television conditions and it is necessary to bear them carefully in mind. The main problem is to obtain sufficient selectivity in the vision channel to prevent the sound signal from interfering with the picture. It is, of course, one which has been present for years in receivers designed for the London station and for this a satisfactory solution has been found. The problem has been intensified in receivers designed for all other stations, however, by the adoption for them of vestigial-sideband transmission. This has made the solution appropriate to London inapplicable.

### Television Standards

In all cases the sound-channel carrier lies 3.5 Mc/s below the vision-channel carrier, which carries modulation frequencies up to some 2.75 Mc/s. For the reception of London, where both sidebands are transmitted, there are three possibilities. Double-sideband reception can be used, in which case a vision-channel bandwidth of some 6 Mc/s at —6 db is needed and the sound-channel comes only 0.5 Mc/s below the lower edge of the vision pass-band. Alternatively, vestigial-sideband reception can be adopted with a 6-db bandwidth of only 3 Mc/s. The sidebands selected can be the upper or the lower. In the latter case, the sound channel again comes only 0.5 Mc/s below the vision pass-band, but if the upper sidebands are selected, the channel separation becomes 3.5 Mc/s.

Needless to say, this last case is a favourite with receiver designers and it is not difficult to obtain adequate discrimination between the two channels. The natural selectivity of the vision-channel intervalve couplings, supplemented by one rejector circuit tuned to the sound signal, usually suffices.

In the case of the new Midland station, and in all future stations, vestigial-sideband transmission is adopted and it is the lower-frequency sidebands, lying between the vision and sound carriers, which are employed. The receiver designer now has no choice but to adopt a 3-Mc/s vision-channel bandwidth with a 0.5-Mc/s separation from the sound channel.

In addition, he has to look forward to the time when more stations are operating and when the sound channel of another station may be 1.5 Mc/s above the vision carrier. Although such a signal may be weak in many districts it will not always be so. Moreover, it will be capable of producing more severe interference than the sound signal belonging to the wanted station. The direct beat between the carriers of the sound and vision signals of one station is at 3.5 Mc/s. Even if this frequency is transmitted through the vision channel, the resulting pattern on the picture is so

fine that it is not very noticeable. The important interference comes from the modulation on the sound channel.

With the sound signal of another station 1.5 Mc/s above the vision carrier, however, the beat will be 1.5 Mc/s and it will cause a very noticeable pattern on the screen. Because of this, it has been recommended that 50-db attenuation should be provided against the sound signal of an adjacent station, while only 30 db is needed against the sound signal of the wanted station. As a result, the receiver must have a vision-channel response curve somewhat as sketched in Fig. 1. The frequency scale has signs correspond ng to the case of an i.f. amp'ifier in a superheterodyne having the local oscillator frequency above the signal frequencies. If the oscillator frequency is below the signal and for a t.r.f. set, the signs must be reversed.

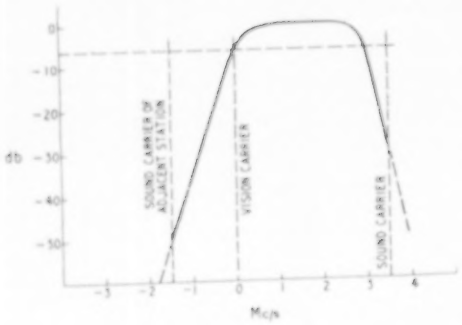


Fig. 1

After this preliminary statement of the selectivity requirements for television, we can turn to our main theme—the relative merits of the superheterodyne and the straight set. In the first place, there is nothing at all in the frequency-changing process which affects selectivity. All the frequency differences existing before the process persist unchanged after it. An interfering signal $\Delta f$ away from the carrier of the wanted signal is still $\Delta f$ away from it no matter to what intermediate frequency the signal is changed. Any difference in selectivity which may exist between superheterodyne and straight set is thus due solely to the difference of operating frequency.

The characteristics of a parallel-tuned circuit, such as that of Fig. 2, are most simply expressed as

$$S = R/Z = \sqrt{[1 + Q^2(1 - x^2)^2/x^2]} \quad \ldots \quad (1)$$

where $Q^2 = CR^2/L$, $x = \omega\sqrt{LC} = \omega/\omega_0 = f/f_0$, $f_0 = 1/(2\pi\sqrt{LC})$, and $S$ is the ratio of the resonant impedance to the impedance away from resonance.

If we write $f = f_0 + \Delta f$ and $\Delta f \ll f_0$ we get the approximate relation

$$S^2 \approx 1 + 4Q^2\Delta f^2/f_0^2 \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

An alternative, and often more useful, result is obtainable. There are two frequencies $f_1 < f_0$ and $f_2 > f_0$ for which the values of $S$ are equal. If $n = f_2 - f_1$

$$S^2 = 1 + Q^2n^2/f_0^2 \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

exactly. If $f_0$ were $(f_1 + f_2)/2$, $n$ would be $2\Delta f$ and equations (2) and (3) would be the same. However, $f_0 = \sqrt{(f_1f_2)}$ and there is some difference between them. Now $Q^2 = CR^2/L$ and $f_0^2 = 1/4\pi^2LC$; therefore $Q/f_0 = 2\pi CR$. The equations can thus be written

$$S^2 \approx 1 + 16\pi^2C^2R^2\Delta f^2 \quad \ldots \quad \ldots \quad \ldots \quad (2a)$$

$$S^2 = 1 + 4\pi^2C^2R^2n^2 \quad \ldots \quad \ldots \quad \ldots \quad (3a)$$

At any level of response $S$ the bandwidth $n = f_2 - f_1$ is quite independent of the resonance frequency and dependent only on the $CR$ product. The same th ng is true of $\Delta f$ as long as $\Delta f \ll f_0$. When it is not, the resonance curve is, in effect, tilted slightly and for equal values of $\Delta f$ the response is greater on one side of resonance than it is on the other. To put it differently $f_2 - f_0 > f_0 - f_1$.

Although we have examined here only a single resonant circuit the results with combinations of circuits are much the same. Cascades of single circuits all resonant to the same frequency obviously are equally independent of the resonance frequency. Stagger-tuned single circuits in cascade are also independent of the mid-band frequency, although this is less obvious, and coupled circuits have similar characteristics. What is true of the single circuit can be taken as being generally true of more complex networks, although there may be minor discrepancies in particular cases when $\Delta f/f_0$ is large.

For the reception of a given signal a certain minimum bandwidth $n_1$ is needed and it is often specified for an edge response of $-6$ db, for which $S^2 = S_1^2 = 4$. For selectivity some other bandwidth $n_2$ is required for which the response is $S_2$. From Equ. (3a) we have

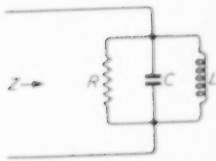


Fig. 2

$$n_2 = \frac{\sqrt{S_2^2 - 1}}{2\pi CR} \quad \text{and} \quad n_1 = \frac{\sqrt{S_1^2 - 1}}{2\pi CR}$$

hence the difference of the bandwidths, $n_2 - n_1$ is constant and quite independent of $f_0$. With any given form of amplifier no control over the selectivity is obtainable by any change of $f_0$ so long as

the pass-band is not greater than that needed for the signal.

In view of this, how did the tradition of the superior selectivity of the superheterodyne arise and why is it still normally used in cases where selectivity is all important?

The answer is that in many common circumstances it is impracticable to make the pass-band only just wide enough for the transmission of the desired signals when the circuits are operating at the frequency of that signal. Equ. (3a) may be misleading here, for it gives the impression that the bandwidth can be reduced to any desired extent merely by increasing $C$ but, in practice, $R$ is likely to vary with an alteration of $C$. Equ. (3) gives a truer picture and at a given frequency bandwidth depends only on $Q$. At any frequency it is proportional to $f_0/Q$, but there is no simple and precise relation between $f_0$ and $Q$. It is found, however, that for coils of a given physical size and type $Q$ is roughly proportional to $\sqrt{f_0}$. The bandwidth is thus roughly proportional to $\sqrt{f_0}$. Consequently for a given bandwidth requirement, there is always some frequency above which it is impossible, impracticable or uneconomical to obtain circuits of high enough $Q$ to meet the requirement.

## Receiver Type

Comparing t.r.f. and superheterodyne receivers having the same form and number of circuits affecting selectivity, the use of the latter is advantageous only when it is impracticable to reduce the bandwidth of the t.r.f. set to the minimum necessary for passing the signal. For the reception of sound signals, and generally whenever $n/f_0$ is very small, the superheterodyne is advantageous. Even with the bandwidth of 2—20 Mc/s needed in radar $f_0$ is so high that the superheterodyne would probably have to be used on selectivity grounds even if it were not already necessary for amplification.

Even in cases where adequate selectivity is possible at signal frequency, the use of the superheterodyne may still be necessary. When a receiver must be tunable over a band for station selection, there is a limit to the number of circuits which can be economically provided at signal frequency. This limit is usually around two or three only. Selectivity may demand the use of a dozen or more circuits, and the proper course is obviously to put most of them at the fixed frequency of the i.f. amplifier.

In the television case, it is easy to secure the minimum bandwidth in the vision channel at signal frequency and up to the present there has been no call for tunable receivers. On the sound side the minimum bandwidth for the transmission of the signal cannot be obtained at signal frequency, but this is not important for, in spite of this, adequate selectivity can still be secured.

It would appear, therefore, that in television the superheterodyne would offer no advantage over the t.r.f. set. This is quite true so long as the receiver does not include rejector circuits for sound-channel rejection or when such rejectors can have a considerable bandwidth. This condition exists in vestigial-sideband reception of the upper sidebands of the London transmitter and then there is little or nothing to choose between the two forms of receiver.

For other stations, however, vestigial-sideband reception of the lower sidebands is necessary. Rejectors of narrow bandwidth are then needed to prevent interference from the sound channel while leaving the vision-channel bandwidth unimpaired. If the required rejector bandwidth is less than can be obtained at signal frequency the superheterodyne has the advantage.

The choice between the two thus hinges not at all upon the main characteristics of the vision channel but entirely upon whether or not it is feasible to obtain rejector circuits of high-enough $Q$ at signal frequency. In connection with rejector bandwidth it should be pointed out that, if it is attainable, it is possible to use a narrower bandwidth in the t.r.f. set than in the superheterodyne because in the latter some allowance must be made for frequency drift of the local oscillator.

We should not like to have to forecast what the eventual decision of designers will be, assuming no extraneous matters enter, for, although it is very hard to obtain a high-enough $Q$ for proper rejector operation at frequencies around 60 Mc/s there are alternatives to the simple rejector. Some bridged-T networks, for instance, enable the equivalent of a rejector with infinite $Q$ to be obtained. The decision will probably be made on extraneous grounds, however, such as the need for a tunable receiver and if so, it will almost certainly be in favour of the superheterodyne.

W. T. C.

# SYNTHESIS OF WIDEBAND TWO-PHASE NETWORKS

By H. J. Orchard, B.Sc.

**SUMMARY.**—Hitherto, networks providing a two-phase supply from a single-phase supply over a wide band of frequencies have been designed empirically. A synthesis technique is now available which gives exact design formulae both for the network components and also for the relations between the design parameters. It is shown that the most general circuit meeting the requirement is essentially a pair of all-pass networks. By utilizing elliptic functions, such networks can be designed to have a Tchebichef approximation to the ideal requirement: this represents the most efficient condition. Simple computing schemes, design curves and a numerical example are included.

## 1. Introduction

IN the design of single-sideband modulator circuits and various other pieces of equipment, pairs of transmission networks are called for having the property that the two members of a pair shall give equal attenuations and have phase shifts differing from one another by $\pi/2$ radians over a wide band of frequencies.

Several publications [1, 2, 3] have been made concerning the theory of such network pairs and various design methods have been proposed, some graphical and some analytical. However, they all proceed, essentially, by analysing some special configuration and trying to adjust its parameters empirically to meet the requirements. This technique is time consuming and, moreover, the methods give no indication of the best performance of which the networks are capable for a given complexity.

Recently it has been found possible to place the whole of the theory on a completely analytical basis which yields both exact design formulae and also clear-cut information as to the theoretical limitations of any possible networks which might be capable of meeting the requirements.

## 2. General Theory

The precise requirement on the network pair is illustrated in Fig. 1. From one harmonic voltage, $V_0$ two more harmonic voltages, $V_1$ and $V_2$ must be derived, such that $V_1$ and $V_2$ are equal in magnitude but different in phase from one another by $\pi/2$ radians. The system may be regarded as a device for producing a two-phase supply from a single-phase supply. However, it is required to operate, not merely at one frequency, but equally well over a specified band of frequencies possibly many octaves wide. It is in this aspect of the problem that all the difficulty lies.

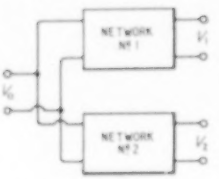


Fig. 1. *Basic phase-splitting circuit.*

Now the phase shift which occurs in any network may, following the nomenclature introduced by Bode[4], be considered as being composed of the minimum phase shift which can accompany the attenuation of the network, and which is uniquely defined by it, together with the phase shift of an all-pass network or phase-equalizer network which may be 'concealed' in the network. If this latter part is zero the network is a minimum-phase network. Thus:

(Total phase shift) = (Minimum phase shift) + (all-pass phase shift).

But both network No. 1 and network No. 2 in Fig. 1 must have equal attenuations at all frequencies in order that $V_1$ and $V_2$ shall be equal in magnitude. Hence the minimum phase shifts for the two networks must be identical. Therefore if $V_1$ and $V_2$ are to be equal for all frequencies they can differ in phase only by an amount which is the difference in phase shift between two all-pass networks. It is pointless to consider, consequently, the case where the two networks, No. 1 and No. 2, are other than all-pass networks, for it would imply that superfluous components existed in the networks, producing identical extra phase shift.

The problem thus resolves itself into one of finding two all-pass networks which have phase-shifts differing from one another by $\pi/2$ radians over a prescribed band of frequencies.

## 3. Properties of All-pass Networks

The most general all-pass network is a symmetrical lattice whose arms are inverse reactances; i.e., the product of their impedances is equal to the square of a constant resistance. Such networks are characterized by the fact that they have a resistive characteristic impedance, independent of frequency and zero attenuation for all real frequencies. However, at real frequencies they do introduce a phase-shift and, within some restrictions, this may be given a

MS accepted by the Editor, August 1949

wide variety of shapes by altering the arm reactances in the lattice. One restriction is that the slope of the phase-shift v. frequency curve must be positive.

The simplest all-pass network is shown in Fig. 2(a) and the one next in complexity in Fig. 2(b): these are referred to as first-degree and second-degree networks respectively. Every all-pass network of higher degree can be replaced by a tandem connection of suitable first- and second-degree networks and such a tandem connection is, from the practical point of view, usually preferable.

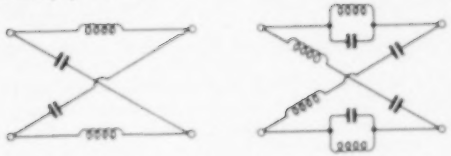

Fig. 2. *All-pass networks.*

The behaviour of an all-pass network, whatever its complexity, is most simply described in terms of its transfer impedance as defined in Fig. 3. If $R_0$ be the characteristic impedance of the network and $Z_T$ the transfer impedance,

$$\text{then } Z_T = v/i$$

For design purposes it is most convenient to normalize the impedances of the network, taking $R_0 = 1$: this is finally changed to some suitable value at the end of the design. The transfer impedance is a rational function, with real coefficients, of the variable $p = j\omega$. The zeros of $Z_T$, which are identical with the transient modes of the terminated network, must, from considerations of stability, lie in the left half of the complex $p$-plane. In order that the impedance shall be appropriate to an all-pass network its modulus must be unity for all real frequencies,

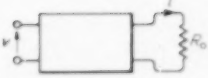

Fig. 3. *Circuit for defining transfer impedance.*

and this requirement forces the poles of $Z_T$ to be equal in number to, and the negative of, the zeros. Hence all the poles lie in the right half of the complex $p$-plane and the transfer impedance may be written as

$$Z_T = \frac{A + pB}{A - pB} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

where $A$ and $B$ are even polynominals in $p$. It will be seen that, at real frequencies, the denominator polynomial in Equ.(1) is the complex conjugate of the numerator, and this forces the requirement of unit modulus on to $Z_T$.

The pole-zero patterns in the complex-plane appropriate to first- and second-degree networks are shown in Fig. 4(a) and 4(b) respectively, where circles denote zeros and crosses, poles. Due to the reality of the coefficients the poles and zeros must, if complex, occur in conjugate pairs.

The phase shift introduced by the all-pass network at real frequencies is equal to the phase angle of $Z_T$ and consequently is given by

$$\tan \tfrac{1}{2}\beta = \frac{\omega . B(-\omega^2)}{A(-\omega^2)} \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

where $\beta$ is the phase shift in radians. Numerator and denominator polynomials introduce equal contributions to the phase at real frequencies as they are conjugates and so it is necessary to consider only the numerator.

Having thus summarized the properties of all-pass networks and their transfer impedances—from such aspects as are necessary,* the next problem is to see how two transfer impedances can be derived, suitable for all-pass networks, and having $\pi/2$ radians difference between their phase angles.

Fig. 4. *Pole-zero patterns.*

## 4. Derivation of Transfer Impedances

Let network No. 1 have a transfer impedance

$$Z_{T1} = \frac{A_1 + pB_1}{A_1 - pB_1} \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

and network No. 2 have

$$Z_{T2} = \frac{A_2 + pB_2}{A_2 - pB_2} \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

We have

$$\tan \tfrac{1}{2}\beta_1 = \frac{\omega B_1}{A_1}; \tan \tfrac{1}{2}\beta_2 = \frac{\omega B_2}{A_2} \quad \ldots \quad (5)$$

and hence, by the addition theorem,

$$\tan \tfrac{1}{2}(\beta_1 - \beta_2) = \frac{\dfrac{\omega B_1}{A_1} - \dfrac{\omega B_2}{A_2}}{1 + \dfrac{\omega^2 B_1 B_2}{A_1 A_2}}$$

$$= \frac{\omega(A_2 B_1 - A_1 B_2)}{(A_1 A_2 + \omega^2 B_1 B_2)} \quad \ldots \quad \ldots \quad (6)$$

* For further properties see Ref. (4) Chap. 11.

$$\text{Put } P = A_2B_1 - A_1B_2 \\ \text{and } Q = A_1A_2 + \omega^2B_1B_2 \Big\} \quad \ldots \quad \ldots \quad (7)$$

so that

$$\tan \tfrac{1}{2}(\beta_1 - \beta_2) = \frac{\omega P}{Q} \quad \ldots \quad \ldots \quad (8)$$

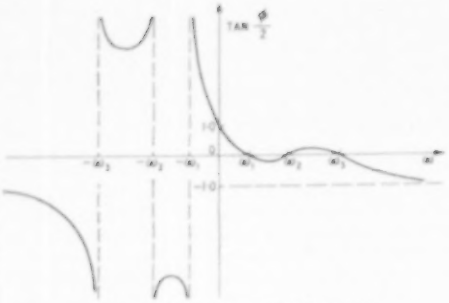


Fig. 5. *Typical curve for* $\tan \phi/2$.

It will be seen that $P$ and $Q$ are even polynomials in $p = j\omega$. Over the prescribed band of frequencies we require $(\beta_1 - \beta_2)$ to approximate, as accurately as possible, to $\pi/2$ and so we are interested, not so much in the magnitude of $(\beta_1 - \beta_2)$, as in the error which is involved; i.e., in the quantity $\left\{\frac{\pi}{2} - (\beta_1 - \beta_2)\right\}$. Forming the tangent of half this error we get

$$\tan\left\{\frac{\pi}{4} - \frac{1}{2}(\beta_1 - \beta_2)\right\} = \frac{1 - \dfrac{\omega P}{Q}}{1 + \dfrac{\omega P}{Q}}$$

$$\text{or } \tan\frac{\phi}{2} = \frac{Q - \omega P}{Q + \omega P} \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

where $\phi = \dfrac{\pi}{2} - (\beta_1 - \beta_2)$.

Thus $\tan\dfrac{\phi}{2}$ is seen to be a rational function whose poles and zeros are the negatives of one another, for the denominator polynomial is derived from the numerator by reversing the sign of $\omega$ (or $p$). Also for $\omega = 0$, $\tan\dfrac{\phi}{2} = 1$. Therefore we can write

$$\tan\frac{\phi}{2} = \prod_{n=1}^{n}\left(\frac{\omega_n - \omega}{\omega_n + \omega}\right) \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

where the $\omega_n$ terms are the zeros of the polynomial $(Q - \omega P)$. As long as the $\omega_n$ terms are either real or occur in conjugate pairs they are quite arbitrary. For the purpose of the present problem it is almost obvious that to get a good approximation over the prescribed band they should all occur at real values of $\omega$ and be distributed over that band. Then the typical form for the function $\tan\dfrac{\phi}{2}$ is as illustrated in Fig. 5.

The problem is now clearly one of finding the best approximation of this rational function over the prescribed band and, under these conditions, of obtaining a numerical relationship between bandwidth, error and the number of $\omega_n$ terms required; i.e., the complexity of the networks.

The most efficient form of approximation is when the $\omega_n$ terms are so chosen that the maximum excursions of $\tan\dfrac{\phi}{2}$ (and hence of $\phi$) from zero are all equal and arranged as shown in Fig. 6. This is the so-called Tchebichef form of approximation and the problem of producing it with a rational function usually involves elliptic functions.

The method of obtaining the Tchebichef approximation is to represent the rational function and its variable by a pair of parametric equations in a manner analogous to that by which, for example, the cycloid is represented by the two equations.

$$y = a(1 - \cos t) \quad x = a(t - \sin t)$$

where the parameter is $t$. For the rational function, however, the parametric equations require not circular but elliptic functions.

Before describing in detail the particular

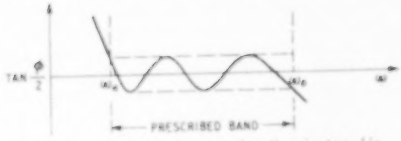


Fig. 6. *Tchebichef approximation to* $\tan \phi/2$.

parametric equations needed here for $\tan\dfrac{\phi}{2}$ and $\omega$ it may be of interest to recall some of the basic properties of the elliptic functions. The elliptic functions are generalizations of the elementary circular functions sin, cos, etc., and stand midway, so to speak, between them and the hyperbolic functions. The most common one is the sn function which is defined by the integral

$$u = \int_0^{\sin(u)} \frac{dt}{\sqrt{(1 - t^2)(1 - k^2t^2)}}$$

the other elliptic functions are defined by similar integrals.

The integral contains the constant $k$ which is called the modulus. (This is an unfortunate

choice of name and it should not be confused with its alternative meaning 'absolute magnitude of.') The elliptic functions are thus seen to be functions of two variables, the main variable $u$, and the modulus $k$. This is denoted symbolicaly in the sn function for example by writing it as sn $(u\ ;\ k)$, and similarly for the others. In general, $k$ may be any complex number, but in most practical applications, as here, it is restricted to be real and lie between 0 and 1.

When $k = 0$ the functions degenerate into circular functions, and when $k = 1$, into hyperbolic functions. For these two special cases the sn integral becomes:

$$u = \int_0^{\sin(u)} \frac{dt}{\sqrt{(1 - t^2)}} \text{ and}$$

$$u = \int_0^{\tanh(u)} \frac{dt}{(1 - t^2)} \text{ respectively.}$$

Now the fundamental property of the circular and hyperbolic functions is that they are periodic: the former have a single real period of $\pi$ or $2\pi$; the latter have a single imaginary period of $j\pi$ or $j2\pi$. The elliptic functions combine these two properties and have both a real and an imaginary period.

The real quarter-periods of the circular and elliptic functions are given by

$$\frac{\pi}{2} = \int_0^1 \frac{dt}{\sqrt{(1 - t^2)}} \text{ and}$$

$$K = \int_0^1 \frac{dt}{\sqrt{(1 - t^2)(1 - k^2t^2)}} \text{ respectively,}$$

while the imaginary quarter-periods of the hyperbolic and elliptic functions are given by

$$j\frac{\pi}{2} = \int_0^{j\infty} \frac{dt}{(1 - t^2)} \text{ and}$$

$$jK' = \int_0^{j\infty} \frac{dt}{\sqrt{(1 - t^2)(1 - k^2t^2)}} \text{ respectively,}$$

or, with a slight change of the variable of integration, $t = j\dfrac{x}{\sqrt{1 - x^2}}$

$$\frac{\pi}{2} = \int_0^1 \frac{dx}{\sqrt{(1 - x^2)}} \text{ and}$$

$$K' = \int_0^1 \frac{dx}{\sqrt{(1 - x^2)(1 - k'^2x^2)}}$$

$k' = \sqrt{1 - k^2}$ is called the complementary modulus.

Due to the double periodicity of the elliptic functions, the complex plane of its variable $u$ is divided up into an infinte lattice of rectangles which are such that, at geometrically similar points in all the rectangles, the function has the same value; these rectangles are called the period rectangles. The sides of the rectangles are $4K$ (or $2K$) and $j4K'$ (or $j2K'$), depending upon the particular function.

Another way of looking at this property is to consider that the complete complex plane of the elliptic function is mapped conformally on to each one of the period rectangles in the complex plane of its variable. This happens also with the circular and hyperbolic functions except that the rectangles degenerate into infinite strips parallel to either the imaginary axis or the real axis respectively.

When the elliptic functions are used to represent parametrically a rational function, say $y = R(x)$, with Tchebichef behaviour, both $y$ and $x$ are mapped with different elliptic functions on to rectangles in the complex plane of the parameter. By using different moduli for the two elliptic functions and by having suitable multiplying constants on the parameter, the rectangle for $x$ is arranged to be some simple integral multiple of the rectangle for y and this guarantees that a *rational* function will be generated.

The two elliptic functions required in the present application are the cd function and the dn function. The cd function is related to the sn function as the cos function is related to the sin function; i.e., they are shifted relative to one another by a real quarter-period ($\pi/2$ or $K$). For real values of the variable the cd and the cos functions behave in very much the same way, oscillating between $\pm 1.0$ and having real periods of $4K$ and $2\pi$ respectively.

The dn function is related to the sn function by $\mathrm{dn}(u) = \sqrt{(1 - k^2\mathrm{sn}^2(u))}$; when $u = 0$, $\mathrm{dn}(u) = 1$ and when $u = K$, $\mathrm{dn}(u) = k'$; it has a real period of $2K$. There is no circular equivalent to the dn function but it degenerates into sech $(u)$ for $k = 1$ and for real values of the variable its behaviour, between $-K$ and $+K$, is somewhat similar to that of sech $(u)$ between $-\infty$ and $+\infty$.

The exact parametric equations are

$$\tan\frac{\phi}{2} = \sqrt{k_1}\,\mathrm{cd}\left[2n \cdot \frac{K_1}{K} \cdot u\ ;\ k_1\right] \quad \ldots \quad (11a)$$

$$\omega = \frac{1}{\sqrt{k'}} \cdot dn\left[u\ ;\ k\right] \quad \ldots \quad \ldots \quad \ldots \quad (11b)$$

With this choice of functions the prescribed band of frequencies is normalized so that the

geometric mean of the end frequencies occurs at $\omega = 1$. $K$ and $K_1$ are the real quarter-periods of elliptic functions of moduli $k$ and $k_1$. $n$ is an additional constant, always integral, which specifies the degree of the rational function being generated: its physical significance will appear later.

$k' = \sqrt{1 - k^2}$ is defined as the ratio of the lower to the upper end frequency of the prescribed band. Thus, in Fig. 6

$$k' = \frac{\omega_a}{\omega_b} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (12)$$

$k_1$ is chosen relative to $k$ so that their corresponding modular constants $q_1$ and $q$ (defined as $q = \exp\left(-\pi \frac{K'}{K}\right)$ etc.) are related by

$$q_1 = q^n \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (13)$$

This relationship, together with the choice of the multiplier $2n \frac{K_1}{K}$ in Equ. (11a), arranges that the two rectangles in the $u$-plane, on to which the $\omega$-plane and the $\tan \frac{\phi}{2}$-plane respectively are mapped, have the same length of side parallel to the imaginary axis but have their sides which are parallel to the real axis, in a ratio of $n$ to 1.

The portion of the curve of $\tan \frac{\phi}{2}$ between $\omega_a$ and $\omega_b$ in Fig. 6 is generated when the parameter $u$ is real. As $u$ increases from 0 to $K$, $\omega$ decreases, by virtue of Equ. (11b) from $\frac{1}{\sqrt{k'}}$ to $\sqrt{k'}$ while at the same time the argument of the cd function in Equ. (11a) increases from 0 to $2n.K_1$ so causing $\tan \frac{\phi}{2}$ to oscillate between $\pm \sqrt{k_1}$ $n$ times. This gives rise to the Tchebichef behaviour. The remainder of the curve is generated when $u$ is complex and travels round the rectangular contour with vertices $K$, $K + j2K'$, $j2K'$, 0.

Having now defined a suitable functional form for $\tan \frac{\phi}{2}$ by means of Equs (11a) and (11b) it is next necessary to find some means of evaluating the poles and zeros of the two transfer impedances. As the poles are the negative of the zeros it will be sufficient to find the zeros, or the poles.

From the definitions of $Q$ and $P$ it will be seen that

$$Q + j\omega P = (A_1A_2 + \omega^2 B_1B_2) + j\omega\,(A_2B_1 - A_1B_2) = (A_1 + j\omega B_1)(A_2 - j\omega B_2) = (A_1 + pB_1)(A_2 - pB_2) \quad \ldots \quad \ldots \quad (14)$$

Now the zeros of $(A_1 + pB_1)$ are the zeros of $Z_{T1}$ and lie in the left half of the $p$-plane while the zeros of $(A_2 - pB_2)$ are the poles of $Z_{T2}$ and so lie in the right half of the $p$-plane. This gives a clue to the means of finding the required poles and zeros: the zeros of $(Q + j\omega P)$ must be found and then expressed in terms of the variable $p = j\omega$, whereupon the zeros of $Z_{T1}$ are those which lie in the left half $p$-plane and the poles of $Z_{T2}$ those which lie in the right half $p$-plane.

The zeros of $(Q + j\omega P)$ are the roots of the equation

$$Q + j\omega P = 0$$

$$\text{or} \quad \frac{\omega P}{Q} = j \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (15)$$

and hence the roots of

$$\frac{1 - \omega P/Q}{1 + \omega P/Q} = \frac{1 - j}{1 + j}$$

$$\text{i.e.,} \quad \frac{Q - \omega P}{Q + \omega P} = -j \quad \ldots \quad \ldots \quad \ldots \quad (16)$$

From Equ. (11a) this is equivalent to

$$\sqrt{k_1}\,\mathrm{cd}\left[2n \frac{K_1}{K} u\,;\, k_1\right] = -j \quad \ldots \quad \ldots \quad (17)$$

The parameter values which satisfy Equ. (17) are given by

$$2n \frac{K_1}{K} u = (4\sigma + 1)K_1 + j\tfrac{1}{2}K_1' \quad \ldots \quad (18)$$

[$\sigma$ = any integer]

By using the relationship between the various quarter-periods, namely

$$4n \frac{K'}{K} = \frac{K_1'}{K_1} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (19)$$

as imposed by the requirement of Equ. (13), Equ. (18) may be written as

$$u = \frac{4\sigma + 1}{2n} \cdot K + jK' \quad \ldots \quad \ldots \quad \ldots \quad (20)$$

Substituting this value for $u$ into Equ. (11b) leads immediately to the required values of $\omega$. Thus the zeros of $(Q + j\omega P)$ are given by

$$\omega = \frac{1}{\sqrt{k'}} \cdot \mathrm{dn}\left[\frac{4\sigma + 1}{2n} \cdot K + jK'\,;\, k\right] \quad \ldots \quad (21)$$

By a slight transformation of the elliptic function this becomes

$$\omega = -\frac{j}{\sqrt{k'}} \cdot \frac{\mathrm{cn}\left[\dfrac{4\sigma + 1}{2n} \cdot K\,;\, k\right]}{\mathrm{sn}\left[\dfrac{4\sigma + 1}{2n} \cdot K\,;\, k\right]}$$

$$\text{or} \quad p = \frac{1}{\sqrt{k'}} \cdot \frac{\mathrm{cn}\left[\dfrac{4\sigma + 1}{2n} \cdot K\,;\, k\right]}{\mathrm{sn}\left[\dfrac{4\sigma + 1}{2n} \cdot K\,;\, k\right]} \quad \ldots \quad (22)$$

The function $\frac{\mathrm{cn}\,[u\,;\,k]}{\mathrm{sn}\,[u\,;\,k]}$ is the elliptic analogue of the cotangent function and degenerates into it when $k = 0$. The cotangent function has a real period of $\pi$ (not $2\pi$) and similarly the function $\frac{\mathrm{cn}\,[u\,;\,k]}{\mathrm{sn}\,[u\,;\,k]}$ has a real period of $2K$, $K$ being analogous to $\pi/2$. For real values of the variable the two functions behave in very much the same way; they are both infinite at $u = 0$ and $u = 2K$ (or $\pi$) and zero at $u = K$ (or $\pi/2$); positive from 0 to K and negative from $K$ to $2K$. The general shape of the function is shown in Fig. 7; altering the value of $k$ alters the ordinates whilst retaining the main characteristics.

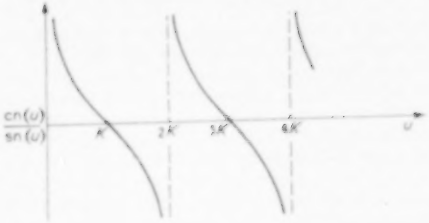


Fig. 7. *Typical curve for* $cn(u)/sn(u)$.

If, in Equ. (22), the integral variable $\sigma$ is given successively the values 0 to $(n - 1)$ inclusive, a set of $n$ different values of $p$ will be generated. Any other values for $\sigma$ will simply repeat some member of this set, for increasing $\sigma$ by $n$ merely adds a whole period to the variable of the elliptic function.

The main result of the analysis, to be noticed at this point, is that, as the variable in the elliptic function is real for all values of $\sigma$, all the values of $p$ which are generated will also be real. This means that the zeros of $Z_{T1}$ and $Z_{T2}$ are all real and that the all-pass networks could be realized physically as a tandem connection of first-degree networks.

Those values of $p$ which are negative are immediately the zeros of $Z_{T1}$; reversing the sign of the positive ones gives the zeros of $Z_{T2}$. It will be seen, after a little examination, that the zeros of $Z_{T1}$ interlace the zeros of $Z_{T2}$. Altogether, distributed between $Z_{T1}$ and $Z_{T2}$, there will be $n$ zeros and hence in the physical realization there would be $n$ first-degree all-pass networks required. The network pair, considered as a unit, may thus be said to be 'of degree $n$.'

This completes the analysis necessary for the derivation of the two transfer impedances. In Sections 6 and 7 will be given methods of computing the elliptic functions of Equ. (22) and details of networks having the transfer impedances so specified. The next problem to be solved is that of finding the relationship between bandwidth, tolerance and network complexity; i.e., between $k'$, $\phi$ and $n$.

## 5. Performance Relationships

Equ. (11a) shows that

$$\tan\frac{\phi}{2} = \sqrt{k_1}\,\mathrm{cd}\left[2n\frac{K_1}{K}u\,;\,k_1\right]$$

The maximum excursion of $\phi$ away from zero coincides with that of $\tan\frac{\phi}{2}$ away from zero. Over the band of approximation $u$ is real and the maximum value of the cd function is $\pm 1$. Therefore the maximum value of $\tan\frac{\phi}{2}$ is $\pm\sqrt{k_1}$.

Denoting the greatest value of $\phi$ by $\phi_m$ it follows that

$$\tan\frac{\phi_m}{2} = \sqrt{k_1} \quad \ldots\ \ldots\ \ldots\ \ldots \quad (23)$$

Normally $k_1 \ll 1$ and then to an extremely good approximation

$$k_1^2 \approx 16q_1 \quad \ldots\ \ldots\ \ldots\ \ldots \quad (24)$$

Using this relationship together with Equ. (13) yields the required form of

$$\tan\frac{\phi_m}{2} = 2q^n \quad \ldots\ \ldots\ \ldots\ \ldots \quad (25)$$

Equ. (25) relates the maximum error in the phase difference $\phi_m$ to the number $n$ of elementary all-pass networks required, and the band of approximation $k'$ which is uniquely related to $q$ by the pair of equations

$$q = \epsilon + 2\epsilon^5 + 15\epsilon^9 + 150\epsilon^{13} + 1707\epsilon^{17} + 20910\epsilon^{21} + \ldots\ \ldots\ \ldots\ \ldots \quad (26a)$$

$$\text{with} \quad \epsilon = \tfrac{1}{2}\cdot\frac{1-\sqrt{k'}}{1+\sqrt{k'}} \quad \ldots\ \ldots\ \ldots \quad (26b)$$

$$\text{or} \quad \sqrt{k'} = \frac{1 - 2q + 2q^4 - 2q^9 + 2q^{16} - \ldots}{1 + 2q + 2q^4 + 2q^9 + 2q^{16} + \ldots} \quad (27)$$

Equs. (25) to (27) are sufficient to allow the examination of any design possibility in a matter of minutes. As a general guide, curves of $\phi_m$, in degrees, against $k'$ for several values of $n$ have been prepared and appear in Fig. 8. The various series involved are highly convergent and in normal cases two or three terms will give adequate accuracy.

## 6. Computation of the Elliptic Functions

In a normal design the parameters $k'$ and $\phi_m$ are prescribed from the beginning and the first operation is to find, according to Equ. (25), the smallest value of $n$ which will meet the requirements. The curves of Fig. 8 will in most cases prove adequate for this purpose. It remains

then to find the zeros of $Z_{11}$ and $Z_{12}$ from Equ. (22) and finally two networks having these transfer impedances.

The evaluation of the elliptic functions in Equ. (22) is simplified to some extent by a consideration of the following points.

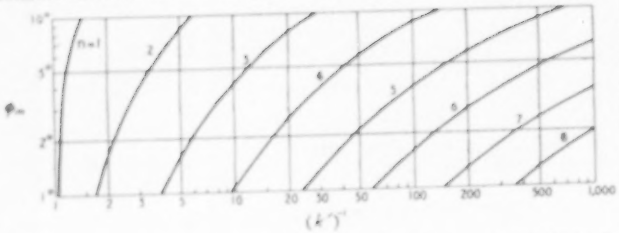


Fig. 8. *Design curves.*

First, the cn function and the sn function are related by the equation

$$\mathrm{sn}^2(u) + \mathrm{cn}^2(u) = 1 \quad \ldots \quad \ldots \quad \ldots \quad (28)$$

and so, if the appropriate sn functions are found, the corresponding cn functions can be computed from this relationship.

Secondly, once it is appreciated that the information provided by Equ. (22) simply implies that the zeros (and poles) of the two transfer impedances interlace one another, it is best to ignore the signs involved and to compute all the quantities as positive numbers; i.e., to compute the poles. They can subsequently be allocated to $Z_{11}$ and $Z_{12}$ from the knowledge that they interlace. It will be found that all the poles are included in the set of numbers defined by

$$\frac{1}{\sqrt{k'}} \cdot \frac{\mathrm{cn}\left[\frac{2\sigma+1}{2n} \cdot K\,;\,k\right]}{\mathrm{sn}\left[\frac{2\sigma+1}{2n} \cdot K\,;\,k\right]}$$

$$(\sigma = 0, 1, \ldots, n-1.)$$

where all the arguments of the sn and cn functions lie in the first quarter-period.

Thirdly, all the poles will be found to be symmetrically disposed on a logarithmic scale around unity: hence it is necessary to compute only half of them, the remainder being found by reciprocation. In addition if $n$ is odd the middle pole will be unity and so needs no computing.

Thus it will be seen that it is necessary to find the values of

$$\mathrm{sn}\left[\frac{2\sigma+1}{2n} \cdot K\,;\,k\right]$$

for values of $\sigma$ from 0 up to the largest number which satisfies the inequality $\sigma < \frac{n-1}{2}$. For instance if $n = 9$ then the $\sigma$ values would be 0, 1, 2, 3.

As far as the evaluation of the functions is concerned tables do exist: however, inasmuch as the elliptic functions are virtually functions of two variables, namely the argument and the modulus, it becomes difficult even in quite bulky tables to provide other than rather coarse gradations of the variables. Hence if tables are used extensive interpolation in two variables must be expected. For this reason the writer has found it preferable to compute the functions directly via a sequence of Landen transformations from the corresponding circular functions. The computing scheme is simple, straightforward and reasonably foolproof, and not the least of its merits is that every number involved is positive. The other practical possibility for computation, that of using the theta functions, is perhaps slightly quicker but it is not so straightforward and great care must be taken with signs.

In its present application the Landen transformation is most conveniently expressed by the equations:

$$\mathrm{sn}\,[\alpha K\,;\,k] = \frac{(1+\lambda)\,\mathrm{sn}\,[\alpha \wedge\,;\,\lambda]}{1+\lambda \cdot \mathrm{sn}^2[\alpha \wedge\,;\,\lambda]} \quad \ldots \quad (29a)$$

where $\lambda = \dfrac{1-k'}{1+k'}$ $\ldots \quad \ldots \quad \ldots \quad \ldots \quad (29b)$

It relates two elliptic functions of different moduli, $k$ and $\lambda$, connected by Equ. (29b), whose arguments are the same fraction of their respective quarter-periods, $\alpha K$ and $\alpha \wedge$. $\lambda$ is smaller than $k$. By repeating the transformation on the modulus indicated by Equ. (29b) only a few times, a point is soon reached where the modulus is vanishingly small. At this point the sn function and the sin function are near enough identical; so the procedure is to replace the sn function by

$$\sin\left[\alpha \cdot \frac{\pi}{2}\right]$$

and, by using Equ. (29a) the appropriate number of times, to retrace the path back to sn $(\alpha K\,;\,k)$. In the present case $\alpha = \dfrac{2\sigma+1}{2n}$. The quantities $\sin\left[\dfrac{2\sigma+1}{2n} \cdot \dfrac{\pi}{2}\right]$ are independent of $k$ and may be tabulated once and for all. Table 1 shows how the scheme is arranged. $k'$ is known and is filled in to start with. Next $\lambda$ is calculated from Equ. (29b) and $\lambda'$ from the relation $\lambda^2 + \lambda'^2 = 1$. This is repeated, a new and smaller modulus being derived from $\lambda'$ by Equ. (29b) and placed

below $\lambda_1$ and so on until a point is reached where it cannot be continued with the number of figures being used because the modulus vanishes and the complementary modulus equals unity.

TABLE 1

| Modulus | Complementary Modulus | 1st Sn Function | 2nd Sn Function |
|---|---|---|---|
| $k$ | $k'$ | | |
| $\lambda_1$ | $\lambda_1'$ | | |
| $\lambda_2$ | $\lambda_2'$ | | |
| $\lambda_3$ | $\lambda_3'$ | | |
| $\lambda_4$ | $\lambda_4'$ | | |
| $\lambda_5$ | $\lambda_5'$ | | |
| $\lambda_6$ | $\lambda_6'$ | | |
| $\lambda_7$ | $\lambda_7'$ | | |
| $\lambda_8$ | $\lambda_8'$ | | |
| $\lambda_9$ | $\lambda_9'$ | | |
| $\lambda_{10}$ | $\lambda_{10}'$ | | |
| Modulus Vanishes | Comp. Mod. $\approx 1.0$ | Sin Function | Sin Function |

Then, alongside this negligible modulus, is placed the sin function of the same fraction of $\pi/2$, as is the required sn function of its quarter-period. By using Equ. (29a) the sn functions of the different moduli in the chain may be computed one after another until finally that for modulus $k$ is reached.

Having found all the sn functions in this way the cn functions are found by Equ. (28) and then, finally, the values of the poles. This yields those poles greater than unity: the remainder, less than unity, are found by reciprocating those computed. To these is added one at unity if $n$ is odd. The poles can now be allocated to $Z_{T1}$ and $Z_{T2}$ by the fact that they interlace.

It is hoped that the numerical example in Section 8 will clarify these points completely.

## 7. Physical Realization of Transfer Impedances

As the poles and zeros of the two transfer impedances are all real it follows, as mentioned in Section 4, that the impedances could be produced by a tandem connection of first-degree networks. In such an event each network is responsible for producing one pole-zero pair in the transfer impedance and the values of the elements, $L$ and $C$, in any one network are given by

$$L = \frac{R_0}{p_r \cdot \omega_0} \qquad C = \frac{1}{p_r \cdot R_0 \omega_0} \quad \ldots \quad \ldots \quad (30)$$

where $R_0$ is the characteristic impedance of the network, $\omega_0$ the geometric mean of the end angular frequencies of the required band and $p_r$ the normalized pole value as given by the computation of Section 6.

All the sections which are joined in tandem must have the same characteristic impedance or, if not, must form an image-matched tandem connection, preferably with the aid of matching attenuators. Also they must be correctly matched at the unparalleled ends.

Despite their simplicity first-degree networks are not the most practicable. This becomes apparent whenever an attempt is made to produce a pair of networks operating over a reasonably wide band of frequencies: then it is found that the networks which are operating over the low-frequency end of the band as 'all-pass' networks are also acting at the higher frequencies as low-pass filters with a cut-off frequency inside the prescribed band. The cause of this is the self-capacitance of the coils, and it being normally impossible to place the self-resonance of the coils in the low-frequency sections outside the complete band.

The remedy for this is to combine pairs of first-degree networks together and realize them as second-degree networks. It is then possible to place across the offending coil an intentional capacitance which can more than absorb the coil self-capacitance. In combining a pair of first-degree networks use may be made of the equivalence of Fig. 9.

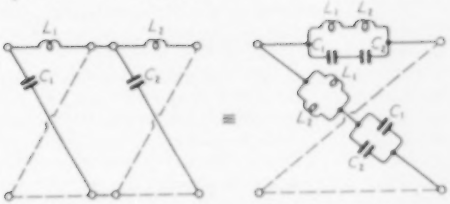


Fig. 9. *Combination of all-pass networks.*

It will be found most advantageous if the pairs which are combined are, first, the ones highest and lowest in frequency; next, the highest and lowest of what remains, and so on until either none or one is left. The impedance of the networks should then be made as low as possible, compatible with the capacitors being not unreasonably large. It may frequently turn out that different impedances are desirable for different networks in the tandem connection. In this case impedance matching attenuators should be fitted between the networks where appropriate.

Where exact equality between the magnitudes of the output voltages is required trouble may be experienced due to dissipation in the networks. This causes an attenuation component, sensibly proportional to $\frac{d\beta}{d\omega}$, in the network as a whole. As $\frac{d\beta}{d\omega}$ is not the same for each arm of the network pair the equality between the voltages tends to be destroyed. One possible remedy

is to use the equivalence indicated in Fig. 10, absorbing a suitable attenuator into each second-degree network and using one of the resistors so produced in each arm to represent the dissipation resistors. It means, that, physically, the element values in the network are changed slightly and one additional resistor is fitted to each arm.

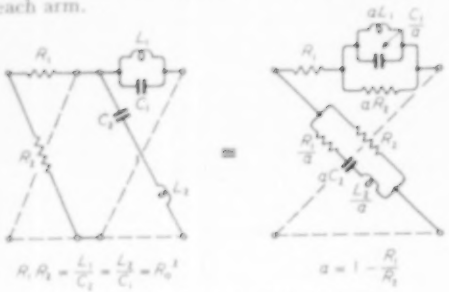


Fig. 10. *Scheme for absorbing dissipation.*

## 8. Numerical Example

The full significance of all the various points dealt with in the preceding sections will, perhaps, be appreciated more easily with the aid of a numerical example. The following design has been chosen with this specific end in view: it is not an example which has occurred in any particular practical problem and it should not be considered necessarily as typical. Usually practical designs are somewhat more stringent and as such are unsuitable for illustrative purposes because they become numerically rather tedious.

Consider the case of a network pair required to operate over the band of frequencies 200 c/s to 4,000 c/s and to have a phase angle tolerance of not more than $\pm 3.0°$. The first problem is to decide upon the degree of the network pair; i.e., the value of $n$. From Fig. 8 it can be seen that for a frequency ratio of 4,000 : 200 = 20 : 1 a network of degree 4 will provide a tolerance of $\pm 2.5°$.

This is checked by means of Equ. (25) to (27) as follows: the complementary modulus, $k'$ is given by $k' = 200/4,000 = 0.05$, and hence $\sqrt{k'} = 0.2236,0680$. From $\sqrt{k'}$ is derived, by Equ. (26b),

$$\epsilon = \tfrac{1}{2} \cdot \frac{1 - 0.2236,0680}{1 + 0.2236,0680} = 0.3172,5600$$

and the sum of the first four terms of the series of Equ. (26a) gives the modular constant $q = 0.3242$. Finally from Equ. (25)

$$\tan \frac{\phi_m}{2} = 2q^4 = 0.0221$$

and so $\phi_m = 2.5°$ thus confirming the figure obtained from the curve.

The next process is to set up the table of Landen transformations: this is shown in Table II.

First, the quantities in the $k$ and $k'$ columns are computed: each quantity in the $k$ column is derived from the $k'$ value in the row above by Equ. (29b) and each value of $k'$ from the adjacent $k$ by the fact that the sum of their squares equals unity. The process terminates after five operations when the value of $k'$ is indistinguishable, to eight significant figures, from unity. Then in the same row and in the columns labelled $\mathrm{sn}[\frac{1}{4}K\,;\,k]$ and $\mathrm{sn}[\frac{3}{4}K\,;\,k]$ are placed the values of $\sin\left[\frac{1}{8}\cdot\frac{\pi}{2}\right]$ and $\sin\left[\frac{3}{8}\cdot\frac{\pi}{2}\right]$ respectively. By using Equ. (29a) and with the value of modulus appropriate to each row, the elliptic function in the row above is obtained. This is repeated until row 1 is filled in.

The Landen transformations give

$$\mathrm{sn}[\tfrac{1}{4}K\,;\,k] = 0.4990,8787$$
$$\mathrm{sn}[\tfrac{3}{4}K\,;\,k] = 0.9284,7394.$$

by Equ. (28)
$$\mathrm{cn}[\tfrac{1}{4}K\,;\,k] = 0.8665,5138$$
$$\mathrm{cn}[\tfrac{3}{4}K\,;\,k] = 0.3713,9755$$

and so
$$\frac{1}{\sqrt{k'}} \frac{\mathrm{cn}[\frac{1}{4}K\,;\,k]}{\mathrm{sn}[\frac{1}{4}K\,;\,k]} = 7.7648,3617$$
$$\frac{1}{\sqrt{k'}} \frac{\mathrm{cn}[\frac{3}{4}K\,;\,k]}{\mathrm{sn}[\frac{3}{4}K\,;\,k]} = 1.7888,9278$$

These are two of the four poles: the other two are the reciprocals of these. Thus the four poles are

1. 7.7648,3617;
2. 1.7888,9278;
3. 0.5590,0499;
4. 0.1287,8571.

Poles No. 1 and No. 3 belong to network No. 1, while poles No. 2 and No. 4 belong to network No. 2. It is now a straightforward matter to compute from Equ. (30) the component values in the elementary first-degree networks and to combine them with the aid of Fig. 9. A suitable value for $R_0$ is 1,200Ω; and $\omega_0$ is given by $2\pi\sqrt{200 \times 4,000} = 5619.85$.

TABLE II

| | $k$ | $k'$ | $\mathrm{sn}[\frac{1}{4}K\,;\,k]$ | $\mathrm{sn}[\frac{3}{4}K\,;\,k]$ |
|---|---|---|---|---|
| 1 | | 0.0500,0000 | 0.4990,8787 | 0.9284,7394 |
| 2 | 0.9987,6191 | 0.4259,1770 | 0.2807,0024 | 0.7093,8393 |
| 3 | 0.4026,0549 | 0.0153,7360 | 0.2034,6320 | 0.5725,0008 |
| 4 | 0.0441,8271 | 0.9990,2347 | 0.1951,8200 | 0.5557,5785 |
| 5 | 0.0004,8850 | 0.9999,9988 | 0.1950,9033 | 0.5555,7025 |
| 6 | 0.0000,0006 | $1.0 - 1.5 \times 10^{-16}$ | 0.1950,9032 | 0.5555,7023 |

The final circuit is shown in Fig. 11.

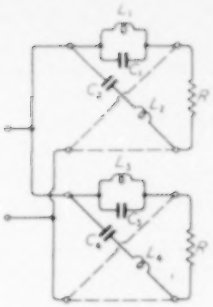


*Fig. 11. Circuit for numerical example.*

*The component values are*

$L_1 = 0.4094$ H
$L_2 = 0.0236$ H
$L_3 = 1.7774$ H
$L_4 = 0.1113$ H
$C_1 = 0.0178\ \mu$F
$C_2 = 0.2844\ \mu$F
$C_3 = 0.0773\ \mu$F
$C_4 = 1.2343\ \mu$F
$R = 1200\Omega$


**Acknowledgments**

This paper is published by permission of the Engineer-in-Chief of the P.O. Engineering Department, in the research laboratories of which the investigation was carried out. The writer is deeply indebted to Dr. J. M. Linke both for his many suggestions in connection with the work and also for his constructive criticism of the manuscript.


REFERENCES


1. U.K. Patent No. 547,001.
2. Dome, R. B.: "Wideband Phase Shift Networks," *Electronics*, Dec. 1946, p. 112.
3. Luck, D. G. C.: "Properties of some Wideband Phase Splitting Networks," *Proc. Inst. Radio Engrs*, Feb. 1949, p. 147.
4. Bode, H. W.: "Network Analysis and Feedback Amplifier Design," D. Van Nostrand, New York, 1945.


# RC-COUPLED POWER STAGE

## *Conditions for Maximum Output*


**By M. G. Scroggie, B.Sc., M.I.E.E.**


FOR drawing the a.c. power output from a valve it is usual to employ a choke or transformer coupling whose d.c. resistance is small and a.c. impedance large compared with either valve or load impedance. By so doing, the device for excluding d.c. from the load can, to a first approximation, be neglected when calculating the power relationships of the stage.

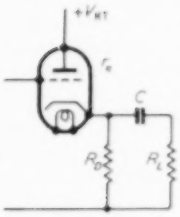


*Fig. 1. The type of coupling considered is shown here as applied to a cathode follower, but the analysis also covers connection in the anode circuit. C is assumed to have negligible impedance at the working frequency, so the effective a.c. load, denoted by $R_A$, is $R_D$ and $R_L$ in parallel.*

There are various familiar expressions, based on this assumption, giving the load resistance for maximum output power as a function of the valve characteristics, depending on the particular conditions laid down; e.g., constant grid drive, constant distortion, etc.

When the valve feed or coupling device is, like the load, a resistance (as for example in Fig. 1) both enter into the problem; and their values for maximum power in the load do not appear to be well known. The conditions for which they will now be derived are those giving maximum 'undistorted' power output of a sine wave; viz., the grid bias and grid drive are so adjusted as to swing the valve exactly between the points at which grid current starts and anode current is cut off. The valve characteristics are assumed linear. The output circuit may be connected on either anode or cathode side of the valve, but in the latter case the ideal conditions which have just been assumed are approximated more closely by reality. (With no negative feedback, distortion usually becomes appreciable before anode-current cut-off is reached). It will be shown that under these conditions the maximum power, $P_{max}$, is obtained in $R_L$ when

$$R_L = r_a$$

and $R_D = \sqrt{2}r_a$

and that when these conditions are fulfilled

$$P_{max} = V^2/93r_a$$

where $r_a$ is given by the nearest linear approximation to the $I_a/V_a$ curve at the threshold of grid current (BE in Fig. 2) and $V$ is the anode supply voltage less the voltage at the point where the $r_a$ line cuts the $V_a$ axis.

Fig. 2 also shows the $R_D$ line, drawn between the point C marking the anode supply voltage, $V_{HT}$, and the $r_a$ line at D. To fulfil the specified conditions, the working point A, must be so placed on the $R_D$ line as to bisect the a.c. load


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line, EF. Since $R_L$, the a.c. load, is $R_D$ and $R_L$ in parallel,

$$R_L = \frac{R_D R_L}{R_D - R_L} \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

Denoting the peak voltage across the load by $v$, and the power output for any values of $R_D$ and $R_L$ by $P$,

$$P = v^2/2R_L \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

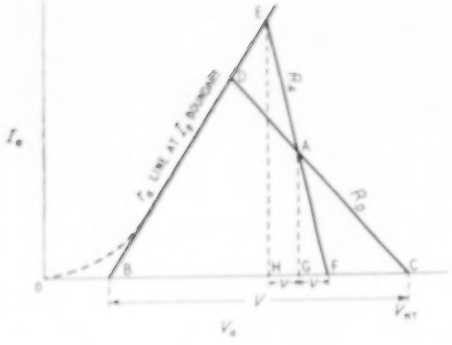



Fig. 2 (above). *A.c. and d.c. load lines on the linearized valve characteristic plane.*

Fig. 3. (right). *Calculated power output (P) vs. $R_L$ as a function of $R_L$ and $R_D$, generalized as regards $r_a$ and V.*

From the geometry of Fig. 2 and the conditions specified

$$v = \frac{HF}{2} = \frac{BH}{2} \cdot \frac{R_L}{r_a}$$

$$\text{So } BH = 2v \cdot \frac{r_a}{R_L} \quad \ldots \quad \ldots \quad (3)$$

$$\text{Also } v = GC \cdot \frac{R_L}{R_D}$$

$$= [V - (BH + v)] \frac{R_L}{R_D}$$

Substituting (3)

$$v = \left[V - \left(2v \frac{r_a}{R_L} + v\right)\right] \frac{R_L}{R_D}$$

$$= \frac{VR_L}{R_L + R_D + 2r_a}$$

$$= \frac{VR_DR_L}{R_DR_L + (R_D + 2r_a)(R_D + R_L)} \quad (4)$$

and, from (2)

$$P = \left[\frac{VR_D}{R_DR_L + (R_D + 2r_a)(R_D + R_L)}\right]^2 \frac{R_L}{2} \quad (5)$$

Differentiating $P$ with respect to $R_L$ and equating to zero,

$$R_L = \frac{R_D(R_D + 2r_a)}{2(R_D + r_a)} \quad \ldots \quad \ldots \quad (6)$$

which gives a maximum for $P$.

Substituting (6) in (5),

$$P = \frac{V^2R_D}{16(R_D^2 + 3r_aR_D + 2r_a^2)} \quad \ldots \quad \ldots \quad (7)$$

Differentiating $P$ with respect to $R_D$ and equating to zero,

$$R_D = \sqrt{2}r_a \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

which is the value giving maximum $P$.

Substituting in (6),

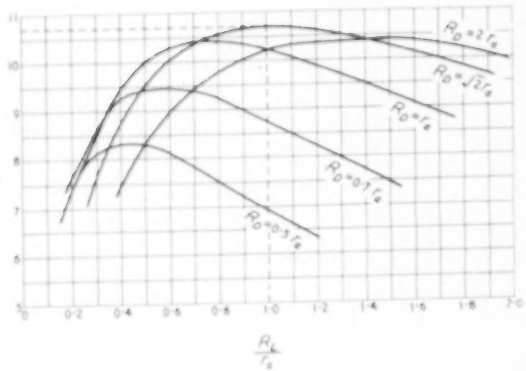
$$R_L = r_a \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

Substituting (8) in (7) to find the maximum power,

$$P_{max} = \frac{V^2}{(48 + 32\sqrt{2})r_a} = \frac{V^2}{93.26r_a} \quad \ldots \quad (10)$$

$$\text{Also } R_A = \frac{r_a}{1 + 1/\sqrt{2}} = 0.586r_a \quad \ldots \quad (11)$$

Using equation (5), the relationship between output power and load resistance is shown in Fig. 3 for several values of $R_D$, from which it can be seen that the optimum values of $R_L$ and $R_D$




are not at all critical. If both are multiplied or divided by any factor not exceeding $\sqrt{2}$, the consequent loss of power does not exceed 3 per cent.

The measured power, using a triode-connected SP61 valve as a cathode follower, with $r_a = 6.6$ kΩ and $V = 210$V, agreed very satisfactorily with the foregoing theory, the calculated $P_{max}$ (72 mW) being obtained close to the point at which distortion of the sinusoidal waveform began to be clearly visible on an oscilloscope.

# TRANSIENT RESPONSE OF A REGULATOR CHAIN

By H. Jefferson, M.A., A.M.I.E.E.

MANY communication systems depend for their satisfactory operation upon the functioning of automatic level regulators, which may be controlled either by the actual signal or by a special pilot signal. One example of such an automatic regulator is the automatic gain-control circuit of a radio receiver, in which the carrier may be regarded as a pilot signal, in contrast to the use of the speech modulation in the constant-volume amplifier which follows the receiver in a single-sideband radio-telephone circuit. Carrier telephone systems on lines and cables, and microwave radio links, may include a number of repeater stations, each incorporating a regulator to maintain the signal at the correct level before transmitting it to the next section.

Level regulators are not instantaneous in action but have a finite operating time. If a change of level occurs at the input to a regulator there is, in consequence, a transient change of level at the output. The effect of this transient on succeeding regulators in the chain is of considerable importance, for it is known that a transient of damped oscillatory character can be set up at points along the chain. These oscillations of level are, of course, associated with oscillations of gain in the amplifiers of which the regulators form a part. Not only are such changes of gain disturbing to the users of the circuit, but should the amplitude of the gain oscillations be excessive a four-wire circuit may oscillate due to excessive loop gain. Should the oscillations of level be excessive, distortion due to overloading may result. The frequency of the oscillations is usually very low, periods of the order of 1 to 10 seconds being common. The frequency is dependent on the characteristics of the level-regulating system.

In the discussion which follows certain assumptions are made regarding the control law of the regulators. These assumptions are chosen to provide a simple mathematical solution, but the characteristics chosen do not deviate too widely from those of practical regulators.

A regulator circuit consists essentially of two components: a variable attenuator in the signal path, and a control or operator circuit bridged across the signal path. The operator circuit measures the level of the signal and produces a voltage or current which is fed to the variable attenuator to adjust the loss. When the regulator is required to maintain a constant output level it is preferable to bridge the operator circuit across the output, thus ensuring a monotonic input-output characteristic. Expander circuits, which belong to the same general class of level-control devices, are best designed with the operator bridged across the input. The two basic units described above are normally supplemented by an amplifier, which provides the required output level.

The portion of the communication system between the output of one level indicator and the output of the next in the chain, which will include a length of cable or a single radio path and one level regulator, will be referred to as a repeater section, and it will be assumed that the regulator is adjusted to provide unit amplification in the normal condition, measured over a repeater section. This is often referred to as zero equivalent.

All gains and losses will be referred to in logarithmic units; which unit is used is unimportant until numerical calculation is involved.

It is assumed that the operator circuit measures the average value of output level, and uses this to control the attenuation. Normally the output level is zero; that is, equal to the reference level for the logarithmic units. (For example, the output level may be 0 db for 1 milliwatt). In this event there is no output from the operator circuit. Averaging may be effected either by rectifying the output voltage and storing charge in a capacitor, or by applying the controlling signal to a thermal element, the temperature of which affects the attenuation.

With these assumptions the characteristics of a repeater section become

$$N_2 = A + N_1 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

$$A = A_0 - k \int_0^t N_2 dt \quad \ldots \quad \ldots \quad (2)$$

where

$N_1$ = input level in logarithmic units
$N_2$ = output level
$A_0$ = 'standard' gain
$A$ = actual gain of the section at time $t$
$k$ = a constant of the system

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Writing

$\int_0^t dt = 1/p$, equations (1) and (2) lead to

$$N_2 = A_0 - \frac{k}{p} N_2 + N_1$$

or

$$\frac{p+k}{p} N_2 = A_0 + N_1 \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

The standard gain $A_0$ is to be zero, so that we have

$$N_2 = \frac{p}{p+k} N_1 \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

If the input level is suddenly increased by one unit (say 1 db) the output level is given by

$$N_2 = \frac{p}{p+k} 1 \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

the solution of which is known to be

$$N_2 = e^{-kt} \qquad t > 0 \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

For a single repeater section, therefore, a sudden increase in input level produces instantaneously an equal increase in output level. The output level then falls exponentially towards the standard value with time constant $1/k$.

For two repeater sections in tandem, we have, for the first,

$$N_2 = \frac{p}{p+k} N_1$$

and for the second section

$$N_3 = \frac{p}{p+k} N_2$$

where $N_3$ is the output from the second section, and $N_2$ the output from the first section is also the input to the second. Clearly, then,

$$N_3 = \left(\frac{p}{p+k}\right)^2 N_1 \quad \ldots \quad \ldots \quad \ldots \quad (7)$$

Considering a step input as before, with $N_1 = 1$

$$N_3 = \left(\frac{p}{p+k}\right)^2 1 \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

$$= p \cdot \frac{p}{(p+k)^2} 1$$

$$= p(te^{-kt})$$

$$= (1 - kt)e^{-kt} \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

A response of this sort has a zero at $t = 1/k$ and a turning point at $t = 2/k$, when the value is $-e^{-2} = -0.135$. It indicates that two repeater sections in tandem, when the input level is raised suddenly, give an output which at first rises by an amount equal to the rise in input. It then falls, passing through the reference level after a time $1/k$ and reaches a minimum at a time $2/k$, after which the level rises monotonically to the reference level. The amplitude of the overswing is 13.5 % of the input step amplitude.

Extending the equations to apply to $n$ identical repeater sections, we have :

$$\left.\begin{aligned} &\text{for the first} \quad N_2 = \frac{p}{p+k} N_1 \quad \ldots \\ &\text{for the second} \quad N_3 = \frac{p}{p+k} N_2 \quad \ldots \\ &\ldots\ldots\ldots\ldots\ldots \\ &\ldots\ldots\ldots\ldots\ldots \\ &\text{for the } n^{\text{th}} \quad N_{n+1} = \frac{p}{p+k} N_n \quad \ldots \end{aligned}\right\} \quad (10)$$

so that

$$N_{n+1} = \left(\frac{p}{p+k}\right)^n N_1 \quad \ldots \quad \ldots \quad (11)$$

Assuming, as before, that $N_1$ is a unit step input, and rearranging (11) slightly :

$$N_{n+1} = p^{n-1} \frac{p}{(p+k)^n} (1) \quad \ldots \quad \ldots \quad (12)$$

The solution of this is known, being

$$N_{n+1} = \frac{d^{n-1}}{dt^{n-1}} \cdot \frac{t^{n-1}}{(n-1)!} e^{-kt} \quad \ldots \quad (13)$$

For a small number of repeater sections this can be evaluated by carrying out the indicated differentiation. A more formal solution, however is obtained by using the Laguerre polynomial.[1]

Then

$$N_{n+1} = \frac{e^{-kt}}{(n-1)!} L_{n-1}(kt) \quad \ldots \quad \ldots \quad (14)$$

The properties of the Laguerre polynomial $L_n(z)$ are known, and tabulated values have been published, although they are not easily accessible. The function has all its zeros, a finite number, between zero and infinity, and they are all real. In consequence, the output level from a regulator chain to which a change of input level is made oscillates a finite number of times about zero level and then decays monotonically to zero level. A system having $n$ repeater sections will have an output response characteristic which crosses the reference level axis $(n-1)$ times. The time scale depends only on $k$, which is a characteristic of the regulators, but the amplitude is independent of $k$ and depends only on the magnitude of the initial disturbance. The actual form can be seen from the figure, which shows the results for up to five repeater sections. It will be noted that the oscillations become more rapid and larger in amplitude as the number of repeater sections is increased.

The condition that $k$ shall be the same for all regulators in the chain is one normally set by the equipment designer. If this limitation is

not made, equations (10) become :

$$N_2 = \frac{p}{p + k_1} N_1 ; N_3 = \frac{p}{p + k_2} N_2 ;$$

$$N_{n+1} = \frac{p}{p + k_n} N_n \quad .. \quad .. \quad (15)$$

so that

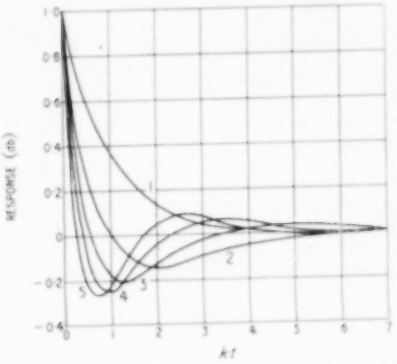
$$N_{n+1} = \frac{p^n}{\prod_{r=1}^{n} (p + k_r)} N_1 \quad .. \quad .. \quad (16)$$

This expression can be expanded into partial fractions, which then give the solution, for a unit step input :

$$N_{n+1} = \sum_{\mu=1}^{n} \frac{(-k_\mu)^{n-1}}{\prod_{r \neq \mu} (k_r - k_\mu)} \exp(-k_\mu t) \quad .. \quad (17)$$

for $t > 0$.

Expressions of this type are encountered in studies of sequences of radioactive disintegrations[2]. As the regulators for which this study was carried out cannot conveniently be constructed to operate with different values of $k$, no further investigation of (17) has been made.




*Relative output level from a system of $(n-1)$ regulated repeaters following a 1-db increase of input level.*

In a practical regulator the integration process is imperfect owing to the presence of leakage. A thermally-controlled system will lose heat by radiation and convection whenever its temperature deviates from the rest condition temperature : a capacitance will lose charge by leakage. If the discharge time constant is $1/m$, the characteristic equation for a single repeater section becomes :

$$N_2 = \frac{p + k}{p + m + k} N_1 \quad .. \quad .. \quad .. \quad (18)$$

For $n$ identical repeater sections, equation (11) becomes :

$$N_{n+1} = \left(\frac{p + m}{p + m + k}\right)^n N_1 \quad .. \quad .. \quad (19)$$

For a unit step change of input level, the output is

$$N_{n+1} = \left(\frac{p + m}{p + m + k}\right)^n \mathbf{1} = \left[p^{n-1} + m(n-1)p^{n-2} + \frac{m^2(n-1)(n-2)}{2!} p^{n-3} + \ldots\ldots + m^{n-1}\right] \frac{p}{(p + k + m)^n} \cdot \mathbf{1} \quad (20)$$

$$= [p^{n-1} + m(n-1)p^{n-2} \ldots + m^{n-1}] \frac{t^{n-1}}{(n-1)!} \cdot e^{-(k+m)t} \quad .. \quad .. \quad (21)$$

$$= \left[\frac{L_{n-1}\{(k+m)t\}}{(n-1)!} + mt \frac{L_{n-2}\{(k+m)t\}}{(n-2)!} + \ldots + m^{n-1} \cdot t^{n-1}\right] e^{-(k+m)t} \quad .. \quad (22)$$

The three solutions derived in this paper provide a way of estimating the effect of the amplitude oscillations which may occur in a practical system incorporating a number of automatic level regulators. They are also of value in showing how the amplitude-oscillation frequency depends on the number of regulators, an effect which is of importance if a final high speed regulator is to be used to eliminate these oscillations. It can be seen from the figure, for example, that after five repeater sections the apparent time constant has been reduced to about one-fifth of that of a single regulator. Deviations from the simple assumptions made do not normally produce any qualitative change in the behaviour of the system.


## Acknowledgment

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## REFERENCES


[1] "Methods of Mathematical Physics," H. and B. S. Jeffreys, Cambridge, 1948. Page 586.

[2] *loc. cit.*, p. 232.

# INDUCED GRID NOISE

By R. L. Bell, B.Sc., Ph.D.

**SUMMARY.**—It is shown that under favourable conditions there is an exact relationship between mean-square noise currents induced at the grid of a triode valve and the 'space-charge' component of input capacitance measurable at that point. In practice this provides an approximate relation useful for estimating the total mean-square induced noise from other measurements.

Techniques of grid-noise measurement are discussed and attention drawn to the occurrence of various inherent errors. Results are presented of measurements of induced grid noise in microwave triodes at 200 Mc/s.

It is concluded that available three-halves power-law treatments are inadequate to explain the induction of shot noise at a control grid, but that values of space-charge admittance measured on the valve together with other low-frequency measurements may be made to yield much closer estimates of induced grid noise in any structure, for moderate transit angles.

## 1. Introduction

IT is widely accepted that the shot-effect fluctuations 'induced' at the control grid are of importance in determining the signal/noise performance of amplifying valves at the higher frequencies. The precise effect of this grid noise, however, is difficult to estimate since reliable methods of calculation are not yet available and measurement of the effects on commercial valves is a prohibitively laborious process. There is evidently required some method of estimating the induced grid noise in terms of other and more readily measurable phenomena.

It is shown in the following that, for small transit angles, the noise induced at the grid is in fact intimately related to the component $C_i$ of input capacitance at that grid traceable to the presence of space charge in the valve. This $C_i$ is *measurable* as the increase in input capacitance of the valve between the cut-off and specified working conditions, provided that both anode and cathode are effectively earthed at the frequency at which the measurement is made.

A random noise current $i(t)$ is conveniently referred to in terms of its 'frequency spectrum' $g(\omega)$ and its 'power spectrum' $s(\omega)$, where $\omega = 2\pi \times$ frequency will be referred to hereafter as 'frequency.' Suppose we are given a definite $i(t)$ in the time interval $(0, T)$, then for that interval

$$g(\omega) = \int_0^T i(t)e^{-j\omega t}\,dt \qquad \ldots \qquad \ldots \quad (1)$$

is defined as its frequency spectrum, and

$$s(\omega) = \lim_{T\to\infty} \frac{2|g(\omega)|^2}{T} \qquad \ldots \quad (2$$

its power spectrum. The factor 2 enters by virtue of the fact that $|g(\omega)|^2$ is an even function of $\omega$ and thus that values of $\omega > 0$ only need be considered.

If we suppose $i(t)$ to flow through unit resistance the time-mean power dissipated will be $i^2(t)$. The quantity

$$\overline{di^2} = s(\omega)\,d\omega$$

involving the power spectrum $s(\omega)$ may be regarded as the contribution to the total power $i^2(t)$ attributable to the Fourier components of $i(t)$ having frequencies between $\omega$ and $\omega + d\omega$.

If for a triode, $s_a(\omega)$ and $s_g(\omega)$ are the power spectra of anode and induced grid noise respectively, $Y_i(\omega)$ $[\approx j\omega C_i]$ the component of grid input admittance due to space charge, and $g_m(\omega)$ the high-frequency mutual conductance, it is shown in Section 3 below that

$$s_g(\omega) = G^2 \left|\frac{Y_i(\omega)}{g_m(\omega)}\right|^2 s_a(\omega) \qquad \ldots \qquad \ldots \quad (3)$$

The value of the factor $G^2$ depends on geometrical and emission conditions inside the valve and is undetermined, except that it is known to lie in the vicinity of unity.

In general, the case $G^2 = 1$ corresponds to the 'ideal valve' in which all electrons are emitted from the cathode with identical normal velocities, and in which the grid is approximated to by a continuous but permeable electrode, field variations parallel to the valve electrodes being eliminated. Since it is not possible to calculate $G^2$ upon the removal of either of these restrictions, it must in effect be assumed that both conditions hold. Attempts at estimating $G^2$ from subsidiary measurements on the valves investigated have not proved very successful, and in practice calculations are made using the approximate value $G^2 = 1$.

## 2. Grid Current and Admittance

We are interested in the short-circuit grid current in a triode in which anode and cathode are connected to a common earth point by leads of negligible impedance. Results may then be derived in terms of the injected currents associated with the 'nodal' method of circuit analysis[1], and the general case of impedances in the leads may be had therefrom by the application of circuit theory. Thus an assumption of zero lead impedances implies no loss of generality.

The relation (2) between the frequency and power spectra of $i(t)$ may be used to deduce the

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following result. Given that a random fluctuation current $i_{12}(t)$ of power spectrum $s_a(\omega)$ flowing in one branch of a (linear) network gives rise to, and is thereby completely correlated with, a fluctuation current $i_{34}(t)$ of spectrum $s_b(\omega)$ in some other branch, the relation between $s_a$ and $s_b$ at a given frequency $\omega$ may be obtained by considering the behaviour of two sinusoidal currents $I_{12}(\omega)$ and $I_{34}(\omega)$, analogous to $i_{12}$ and $i_{34}$ and flowing in the specified branches. One has in fact that

$$\frac{s_b(\omega)}{s_a(\omega)} = \left|\frac{I_{34}(\omega)}{I_{12}(\omega)}\right|^2 \quad .. \quad .. \quad .. \quad (4)$$

as might be expected.

If we suppose that a sinusoidal current $I_{12}(\omega)$ flows in the branch joining nodes 1 and 2 of a valve circuit and gives rise, by any mechanism whatever, to the current $I_{34}$ in another branch, the fundamental assumption of all small-signal valve analyses may be made; viz., that the two currents are linearly related:

$$I_{34}(\omega) = \rho(\omega)\, I_{12}(\omega) \quad .. \quad .. \quad .. \quad (5)$$

where $\rho(\omega)$ is some complex circuit parameter. Suppose now an impedanceless signal generator of voltage $V_{34}(\omega)$ to be applied between terminals 3 and 4. This will give rise to a current

$$I_{12} = Y_T(\omega).\ V_{34}$$

where $Y_T(\omega)$ is the transfer admittance relating $I_{12}$ to $V_{34}$. The resultant current flowing between 3 and 4 will therefore be expressible in the form

$$I_{34} = Y_{oi}\, V_{34} + \rho\, Y_T\, V_{34}$$

the input admittance between 3 and 4 being

$$Y_i = Y_{oi} + \rho Y_T$$

Whence in terms of admittances

$$\rho = \frac{Y_i - Y_{oi}}{Y_T} \quad .. \quad .. \quad .. \quad .. \quad (6)$$

Thus we have for fluctuation power spectra

$$s_b(\omega) = \left|\frac{Y_i - Y_{oi}}{Y_T}\right|^2 s_a(\omega) \quad .. \quad .. \quad (7)$$

where $s_a(\omega)$ is the spectrum of some fluctuation internal to the network, $s_b$ the spectrum of the observable fluctuation due solely to the source $s_a$, $Y_i$ the input admittance at the point of observation and $Y_T$ a transfer admittance from that point to the above internal branch of the network.

Since $\rho$ is finite [$\rho < \infty$], from (6) $Y_{oi}$ is the value of $Y_i$ observable when $Y_T = 0$. In general $Y_T$ will be some function $Y_T(\rho)$ of the 'feedback coefficient' $\rho$ and will be expressible in the form

$$Y_T(\rho) = Y_T(0) + \rho Y_T'(0) + \frac{\rho^2}{2} Y''_T(0) \ldots\ldots$$

Substituting this in (7),

$$\frac{s_b(\omega)}{s_a(\omega)} = \left|\frac{Y_i - Y_{oi}}{Y_T(0)}\right|^2 \left|1 + \rho\,\frac{Y_T'(0)}{Y_T(0)} + \frac{\rho^2}{2}\frac{Y_T''(0)}{Y_T(0)} \ldots\ldots\right|^{-2} \quad (8)$$

The values of the terms $\rho\,\dfrac{Y_T'(0)}{Y_T(0)}$, $\dfrac{\rho^2}{2}\dfrac{Y_T''(0)}{Y_T(0)}$, etc, encountered in the problems to be discussed are such that their contribution to the second bracket is in the nature of a small 'correction' only. Neglecting this effect, we have to a close approximation in the presence of feedback

$$s_b(\omega) = \left|\frac{Y_i - Y_{oi}}{Y_T(0)}\right|^2 s_a(\omega) \quad .. \quad .. \quad (9)$$

Here $Y_T(0)$ is the value of $Y_T$ calculable in the absence of feedback ($\rho = 0$).

In calculating the fluctuation power spectrum appearing at a valve grid, we let $s_b$ and $s_a$ be the power spectra of the short-circuit current fluctuations in the grid-earth and cathode-anode branches respectively. The value $Y_T(0)$ of the transfer admittance $Y_T$ in the absence of feedback to the grid is simply the high-frequency mutual conductance $g_m(\omega)$ of the valve. $Y_i$ is the input admittance at the valve grid under actual operating conditions and $Y_{oi}$ the value of $Y_i$ when $g_m = 0$; i.e., at cut off. The difference $Y_i - Y_{oi}$ may be expressed as an 'active' component of grid input admittance $Y_1$.

Hence, as a close approximation for the grid fluctuation spectrum $s_g(\omega)$, we have

$$s_g(\omega) = \left|\frac{Y_1(\omega)}{g_m(\omega)}\right|^2 s_a(\omega) \quad .. \quad .. \quad (10)$$

where the admittance $Y_1$ itself arises from the effects of feedback to the grid. Examples of such admittances are the 'space charge' component of input capacitance at a grid due to finite electron-transit time, and the input conductance at a grid due to anode and cathode lead-inductance effects.

## 3. The Distributed Case

In this Section we investigate the effect on the expression (10) of a variation of the feedback coefficient $\rho$ to the grid, with say emission energy and trajectory of the electron. The discussion will be relevant only to the case of the 'induced' component of grid noise (i.e., that due to electron transit-time effects) and will clearly *not* apply to external 'circuit' effects such as the lead-inductance feedback grid noise, since all electron groups are treated indistinguishably by external feedback mechanisms.

It is easily shown that for a sufficiently negative grid the potential minimum in a triode will attach itself to the grid wires and approach the cathode only in the regions between, as shown in Fig. 1.

Considering the electrons of emission energy $\lambda$ which finally arrive at the anode from a point $z$ on the cathode of Fig. 1; one of these electrons will cause an anode current which is a func-

tion of $\lambda$, $z$, and time $t$, $i_a(t, \lambda, z)$ and will at the same time induce a grid-current pulse $i_g(t, \lambda, z)$. These two currents, being due to the single event of the transit of a particular electron, will be completely correlated. Taking the Fourier transform (deriving frequency spectra) of these currents gives the two harmonic 'currents' $I_a(\omega, \lambda, z)$ and $I_g(\omega, \lambda, z)$ which are completely defined, one implicitly in terms of the other. The conditions for the application of the admittance relation (10) are thus satisfied by the transit of every electron of this type, and some form of (10) will therefore apply to the grid noise induced by the 'transmitted' type of electron.

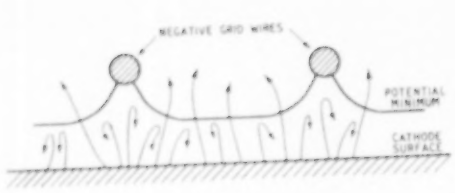



Fig. 1. *Shape of potential minimum and electron trajectories for a negative grid.*

Considering next an electron which fails to cross the potential minimum and returns to the cathode in the region midway between grid wires, it will momentarily effect the space-charge distribution in the triode, giving rise to an anode current $i_a(t, \lambda, z)$. The net effect will induce a grid current $i_g(t, \lambda, z)$ and the two currents will specify each other completely. We are led to the result that the admittance relation applies equally to electrons of this group. Moreover, it is clear that there is no essential difference between the type of reflected electron just discussed and that which is emitted from that part of the cathode immediately under a grid wire, for all of these will have some effect on the anode current.

We have then that for all $\lambda$ and $z$ there is an induced grid current $i_g(t, \lambda, z)$ with a spectrum $I_g(\omega, \lambda, z)$, anode currents $i_a(t, \lambda, z)$ and $I_a(\omega, \lambda, z)$, and that the corresponding anode and grid currents are completely correlated. We may, therefore, attempt an application of (10) to the total induced grid noise.

If $\mu(\lambda, z)\, d\lambda\, dz$ be the rate of emission of electrons of normal component of emission energy in the range $(\lambda, \lambda + d\lambda)$ from a region $(z, z + dz)$ of the cathode, and the frequency spectrum of the grid-current pulse due to one of these electrons be $I_g(\omega, \lambda, z)$, then following the method of a previous paper[2] the power spectrum of the total induced grid noise is given by a repeated application of Campbell's theorem:

$$s_g(\omega) = 2\int d\lambda \int_z \mu(\lambda, z) \left| I_g(\omega, \lambda, z) \right|^2 dz \qquad (11)$$

Clearly, if $Y_1(\omega, \lambda, z)\, d\lambda\, dz$ and $g_m(\omega, \lambda, z)\, d\lambda\, dz$ are respectively the contributions to grid admittance and mutual conductance made by electrons in the above group $d\lambda\, dz$, we can write

$$I_g(\omega, \lambda, z) = \frac{Y_1(\omega, \lambda, z)\ d\lambda\, dz}{g_m(\omega, \lambda, z)\ d\lambda\, dz}\ I_a(\omega, \lambda, z) \qquad (12)$$

where $I_a(\omega, \lambda, z)$ is the frequency spectrum of the anode-current pulse due to an electron of that group. It is to be noted that the angle of emission of the electron (or alternatively transverse emission energy) can be taken into account by a slight extension of the procedure here, and will therefore be ignored.

One has then from (11) and (12) that

$$s_g(\omega) = 2\int_\lambda d\lambda \int_z \mu(\lambda z) \left| \frac{Y_1(\omega,\lambda,z)}{g_m(\omega,\lambda,z)} \right|^2 |I_a(\omega,\lambda,z)|^2 dz \qquad (13)$$

and similarly for the anode noise

$$s_a(\omega) = 2\int_\lambda d\lambda \int_z \mu(\lambda, z) \left| I_a(\omega,\lambda,z) \right|^2 dz \qquad (14)$$

For the 'ideal valve' with single-valued fields and velocities, $Y_1(\omega, \lambda, z)$ and $g_m(\omega, \lambda, z)$ no longer vary with $z$, and only a very narrow band of values of $\lambda$ is possible, hence

$$\left| \frac{Y_1(\omega,\lambda,z)}{g_m(\omega,\lambda,z)} \right|^2 = \frac{\left| \int_z \int_\lambda Y_1(\omega,\lambda,z)\ d\lambda\, dz \right|^2}{\left| \int_z \int_\lambda g_m(\omega,\lambda,z)\ d\lambda\, dz \right|^2} = \left| \frac{Y_1(\omega)}{g_m(\omega)} \right|^2$$

This factor may then be taken outside the integrals of (13) and there follows from (13) and (14) the result (10):

$$s_g(\omega) = \left| \frac{Y_1(\omega)}{g_m(\omega)} \right|^2 \cdot s_a(\omega) \qquad (10)$$

In any other situation we can combine (13) and (14) only by writing, by analogy with (10),

$$s_g(\omega) = G^2 \left| \frac{Y_1(\omega)}{g_m(\omega)} \right|^2 \cdot s_a(\omega) \qquad (15)$$

where the factor $G^2$, which may be greater or less than unity, takes into account the 'distributed correlation' between anode and grid currents, as expressed in the variation over $\lambda$ and $z$ (initial energy and point of emission of the electron) of the factor

$$\left| \frac{Y_1(\omega,\lambda,z)}{g_m(\omega,\lambda,z)} \right|^2 \qquad (16)$$

In many cases of practical interest $G^2$ will not be much different from unity, since the quotient (16) will tend to be independent of $\lambda$ at least, but this is by no means necessary.

## 4. Microscopic Theory

There have already been made [2, 4, 5] various attempts at estimating the noise induced at a control grid in terms of the transit time in an equivalent simple system, such as the 'three-halves power-law.' Unfortunately, however, these treatments have so far made such sweeping assumptions as almost to invalidate subsequent conclusions. For instance, all considerations of the potential minimum and of essential phenomena connected therewith are omitted—it is assumed that all electrons are emitted with the same (zero) velocity, and thus necessarily that all induced electrode currents are similar and are similarly affected by space-charge damping phenomena. This turns out on examination to constitute a serious limitation, for in practice both the induced grid current and the magnitude of space-charge damping effects[6] may vary considerably with emission energy. Further, no account is taken of the fact that the grid is actually at a potential very different from that effective at the grid plane, magnitudes of induced grid currents being dependent on actual grid-cathode potential. Thus in general, no great confidence can be placed in results so far available.

An attempt is being made to develop an approximate 'microscopic' theory of induced grid noise, bearing these points in mind, by considering the effect on the grid of the transit of a particular electron group, and integrating to obtain the accumulated effect. In this article, however, there will be adopted a result of Campbell and others[8] derived by a similar procedure but based on pure three-halves power-law conditions; viz.,

$$s_g(\omega) = \left(\frac{\omega\tau}{5}\right)^2 \cdot 2eI_a\Gamma^2 \quad \ldots \quad \ldots \quad (17)$$

Here $\tau$ is the electron transit time in a three-halves power-law diode, with cathode at the potential minimum and anode at the grid plane, $\omega\tau$ the corresponding 'transit angle' and $2eI_a\Gamma^2$ the anode shot-noise spectrum of the triode.

All of the quantities on the right-hand side of (17) are measurable on the valve, yielding an estimate of the induced grid noise which can immediately be compared with measured values.

## 5. Measurements.—Neutralization of Lead-Inductance Noise

In making measurements on the induced grid noise, it is convenient to reproduce as closely as possible the assumed conditions of Section 2—that the anode and cathode of the triode are connected to a common earth point by leads of very low impedance. At the frequencies involved, however, there remain inevitably residual anode- and cathode-lead inductance effects which make undesirable contributions to fluctuations observed at the grid. The active component of grid input admittance consists of lead-inductance and transit-time admittances respectively:

$$Y_i = Y_L + Y_T$$

the total fluctuation at the grid being given, therefore, by

$$s_g(\omega) = \left|\frac{Y_L + Y_T}{g_m}\right|^2 s_a(\omega) \quad \ldots \quad \ldots \quad (18)$$

(In this Section and the next it is sufficient to assume that the 'admittance relation' applies in its simple form (10) to the induced effect, as to the lead-inductance effects). The active part of the lead-inductance admittance (see Section 2) is given to a second-order accuracy by

$$Y_L = g_m(\omega^2 L_k C_{gk} - \omega^2 L_a C_{ga}) \quad \ldots \quad (19)$$

which is purely conductive. The induced component of grid noise is accounted for by the 'transit-time' component of input admittance

$$Y_T = G_T + jB_T \quad \ldots \quad \ldots \quad \ldots \quad (20)$$

and may therefore be expressed as

$$s_{gT}(\omega) = s_a(\omega)\left\{\frac{G_T^2 + B_T^2}{g_m^2}\right\} \quad \ldots \quad \ldots \quad (21)$$

In the working range of a valve functioning as a low-noise amplifier, the approximate 3/2 power-law theory suggests, and measurements confirm, that

$$B_T^2 \gg G_T^2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (22)$$

Hence to a sufficiently close approximation, one has for the two components of input admittance

$$Y_L = G_L, \qquad Y_T = jB_T \quad \ldots \quad \ldots \quad (23)$$

where $B_T = \omega C_t$ is the 'space-charge component' of input susceptance. The total grid-noise spectrum according to (18) is then

$$s_g(\omega) = \left\{\frac{G_L^2 + B_T^2}{g_m^2}\right\} s_a(\omega) \quad \ldots \quad \ldots \quad (24)$$

and the induced grid noise, from (21) is

$$s_{gT}(\omega) = \frac{B_T^2}{g_m^2} s_a(\omega) \quad \ldots \quad \ldots \quad \ldots \quad (25)$$

which is obtained from (24) by setting $G_L = 0$. This is in turn effected by satisfying in (19) the 'bridge relation'

$$L_a C_{gk} = L_k C_{ga}$$

by introducing extra reactance into the anode lead and *adjusting this for minimum noise output* from the grid; i.e., from (24)

$$\frac{d}{dG_L} s_g(\omega) = \frac{2G_L}{g_m^2} s_a(\omega) = 0$$

therefore, $G_L = 0 \quad \ldots \quad \ldots \quad \ldots \quad (26)$

and $s_g(\omega) = \frac{B_T^2}{g_m^2} \cdot s_a(\omega) \quad \ldots \quad \ldots \quad \ldots \quad (27)$

as required.

It is of interest to note, by way of a check, that substituting for $G_L$ from (19), the mean-square

noise current transferred to the grid circuit from the anode circuit by lead inductance effects alone is given by

$$s_{gL}(\omega) = |\omega^2 L_k C_{gk} - \omega^2 L_a C_{ga}|^2 s_a(\omega)$$

which is the result derived by direct application of circuit theory to the problem.

In order to avoid major difficulties with lead-inductance effects, actual measurements of grid noise were performed on u.h.f. disc-seal triodes, having very low residual lead inductances.

## 6. Measurement Process

The preliminary sketch given in Section 5 above, while indicating the principles of neutralizing to be observed in dealing with the lead-inductance feedback noise, omits all reference to the effects of associated impedance changes on the measuring instrument. The first stage of the amplifier used for the measurements employs a microwave triode in common grid connection, an equivalent input circuit appearing in Fig. 2.

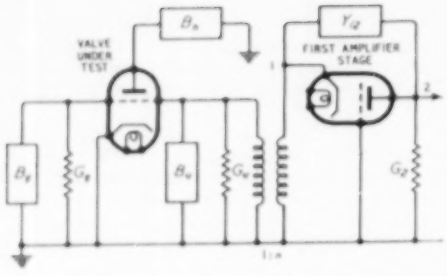



Fig. 2. *Equivalent measurement circuit.*

If $s_g(\omega)$ be the power spectrum of the noise current appearing at the grid of the valve under test, the spectrum of the output from the first stage is given by

$$s_2(\omega) = \left\{\frac{s_g + 4kTG_u}{n^2}\right\}\left|\frac{\Delta_{12}}{\Delta_{22}}\right|^2 + 2eI\,\Gamma^2\left|\frac{\Delta_{22} - \Delta_{12}}{\Delta_{22}}\right|^2 + 4kTG_2 \quad \ldots \quad (28)$$

In this expression, 1 refers to cathode and 2 to anode of the common-grid amplifier,

$$\Delta_{22} = g_m + g_a + Y_{19} + Y_{12}$$
$$\Delta_{12} = g_m + g_a + Y_{12}$$
$$B_1 = B_g + B_u$$
$$G_1 = G_g + G_u$$
$$Y_{19} = \frac{G_1 + jB_1}{n^2}$$

Where $B_g$ is the grid input susceptance of the test valve and $B_u$ is the adjustable grid circuit susceptance, $G_u$ and $G_g$ corresponding conductances, and $n$ a transformation ratio defined later. $g_m$ and $g_a$ are the mutual and anode conductances respectively and $2eI\,\Gamma^2$ the anode shot-effect spectrum of the common-grid amplifier. The quantities $Y_{12}$, the admittance of the feedback path across the amplifier, and $G_2$, its anode-load conductance, may be neglected here, when the input noise power spectrum equivalent to (28) above appears as

$$s_{1\sigma} = s_g + 4kTG_u + 2eI\,\Gamma^2\left\{\frac{G_1^2 + B_1^2}{n^2\,g_c^2}\right\} \quad (29)$$

where $g_c^2 = (g_m + g_a)^2$

The procedure in measurement is to note the quiescent value of the output noise associated with $s_{1\sigma}$ [i.e., corresponding to the value of (29) with $s_g = 0$]; the test valve is then switched on, with the following effects. The input circuit is detuned by the extra space-charge capacitance thrown across it, $G_1$ is slightly varied by the onset of transit-time and lead-inductance damping $G_g$ at the test grid, and a small fluctuation current is injected by the valve into the new impedance. The capacitance change is tuned out, the conductance change partly neutralized (as will appear later) and the net change in output noise noted. The test valve is then turned off, a standard source (noise diode) tuned on, the circuits retuned and the standard injected fluctuations adjusted to cause the same small deflection of the output meter. The amount of the standard fluctuations is noted, and is assumed to be a measure of the test-valve grid fluctuations.

It is to be noted that $s_g$, which we wish to measure, is very much smaller than the rest of the terms on the right-hand side of (29). Thus apart from various rudimentary difficulties, the accuracy of the measurement is particularly sensitive to variations during the measurement process of these 'quiescent' terms, due to slight changes of amplifier gain and to variations of the total input-circuit conductance $G_1$.

The same consideration necessitates also that the amplifier be worked with maximum attainable signal/noise sensitivity, which entails that the impedance viewed from the cathode input terminal shall have a fixed optimum value, in practice very nearly conductive at the mid-band frequency. This optimum conductance is obtained experimentally by adjustment of the transformation ratio $1 : n$ from the grid of the test valve to the input of the amplifier, the value of which will vary according to the value of grid-circuit conductance $G_1$.

The susceptance $B_u$ is adjusted throughout the measurement independently of any other adjustment to give minimum noise output from

the amplifier; i.e.,

$$\frac{d}{dB_a} s_{1\alpha} = \frac{d}{dB_1} s_{1\alpha} = 2eI\,F^2 \cdot \frac{2B_1}{n^2 g_c^2} = 0$$

whence $B_1 = 0$ .. .. .. .. (30)

It is clear from the work of Section 5 that, if we are to have a reasonable estimate of the induced grid noise (the admittance change associated with it is mainly susceptive and can, therefore, be tuned out) conductance changes $G_g$ at the grid involved in switching on the test valve will have to be small compared say with the resistive grid-circuit conductance $G_a$. In calculating their effects a first-order accuracy is obtained by assuming that they are, in fact, of differential order compared with $G_a$. Substituting (30) in (29) we have then that the increase in 'output' noise on switching on the test valve is given in terms of the actual injected noise $s_g$ and the conductance change $dG_1 = G_g$ by

$$s_1 = s_g + 2eI\,F^2 \cdot \frac{2G_1}{n^2 g_c} \cdot G_g \quad .. \quad .. \quad (31)$$

where by (26)

$$s_g(\omega) = \left\{ \frac{(G_L + G_c)^2 + B_T^2}{S^2} \right\} s_a(\omega)$$

$S$ is the test-valve mutual conductance, $s_a(\omega)$ its anode shot effect spectrum and

$$G_g = G_L + G_T$$

the sum of the lead-inductance and transit-time conductances (see Section 5). It is of interest to note at this point that the spectrum of the anode noise in the test valve can be written approximately[6]

$$s_a(\omega) \approx 3 \times 4kT \cdot S$$

the cathode being run at normal temperatures. One has also, for the first amplifying stage

$$2eI\,F^2 \approx 3 \times 4kT \cdot g_c .$$

If then we suppose that by some method we have set $G_L$ (and thus the lead-inductance grid noise) equal to zero, we have from (31) that

$$s_{1\alpha} = 3 \cdot \frac{B_T^2 + G_T^2}{S^2} \cdot 4kTS + 3 \times 4kT\,g_c \cdot \frac{2G_1 G_T}{n^2 g_c^2}$$

i.e.,

$$s_{1\alpha} = 3 \times 4kTS \left\{ \frac{B_T^2 + G_T^2}{S^2} + \frac{G_T}{S} \cdot \frac{2G_1}{n^2 g_c} \right\} \quad (32)$$

The first term in the bracket of (32) represents the induced grid noise in the valve under test; the second indicates an error in the measurement due to transit-time damping of the input circuit by the test valve. The expression $G_1/n^2$ is the admittance of the input circuit as viewed from the amplifier, which is fixed by the noise-factor requirements. If the performance of the first valve is of a high standard (i.e., the value of this optimum source admittance is low) the coefficient $2G_1/n^2 g_c$ of the error term may be quite small. It is clear from (29) that there is nothing to be gained by heavy damping of the input circuit (i.e., increasing $G_a$ and $G_1$) since this will have to be taken up by an adjustment of the transformation ratio $n$, and will only make the measurements more difficult by decreasing the sensitivity of the measuring gear.

If, as in the elementary neutralizing process discussed previously, we decide to vary $G_L$ until the noise output is a minimum, (31) becomes

$$s_{1\alpha} = 3 \times 4kTS \left\{ \frac{B_T^2 + G_L^2}{S^2} + \frac{2G_1}{n^2 g_c} \cdot \frac{G_L + G_T}{S} \right\}$$

.. .. .. .. (33)

Setting $\frac{d}{dG_L} s_{1\alpha} = 0$

gives $\frac{G_L}{S} = -\frac{G_1}{n^2 g_c}$

and (33) becomes

$$s_{1\alpha} = 3 \times 4kTS \left\{ \frac{B_T^2}{S^2} + \frac{2G_1 G_T}{n^2 g_c S} - \frac{G_1^2}{n^4 g_c^2} \right\} \quad (34)$$

The first term in (34) represents quite closely the induced grid noise, the second term is again the positive transit-time damping error, and the third represents a fixed negative error, depending with the $G_T$ error on the conductance ratio $\frac{G_1}{n^2 g_c}$. The two errors may be expected to cancel somewhere in the working range of the valve. This 'minimizing' process is that most conveniently used in practice.

The above analysis is given in order to demonstrate the errors inherent in measurements of small amounts of induced grid noise, and is not applicable to quantitative correction of the results, since for instance it has not taken into account the effects of impedance changes in the first stage on the feedback conditions in the second. These effects are not, in practice, negligible compared with those we have already discussed, but are not calculable with any degree of confidence.

## 7. Subsidiary Measurements

Substituting in (25) for the quantity $B_T$, one has that the induced grid noise spectrum is given by

$$s_g(\omega) = \left| \frac{j\omega C_T}{g_m} \right|^2 \cdot G^2 \cdot s_a(\omega) \quad .. \quad .. \quad (35)$$

where the anode shot noise $s_a(\omega)$ is given at low frequencies by the familiar expression

$$s_a = 2e\,I_a\,F^2 \quad .. \quad .. \quad .. \quad .. \quad (36)$$

To check (35) against grid-noise measurements, we have to measure or to estimate the behaviour with variable electrode potentials of each of the quantities $C_s$, $G^2$, $g_m$, and $s_a$ on the right-hand side.

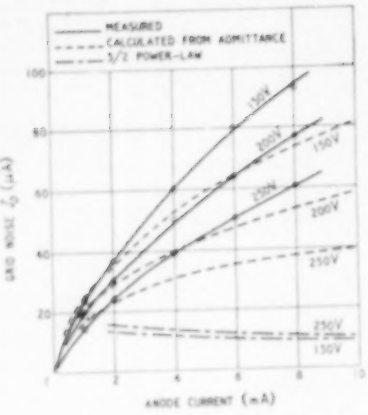



*Fig. 3. Induced grid noise in experimental type E1600, No. 321A, at 200 Mc/s; voltages indicated are those on the anode.*

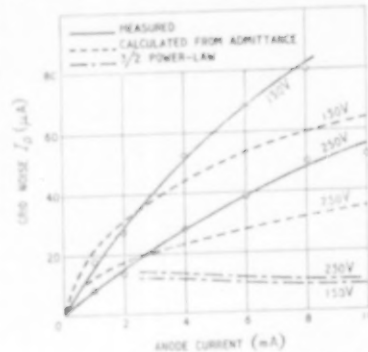



*Fig. 4. Induced grid noise in experimental type E1600, No. 273A, at 200 Mc/s; voltages indicated are those on the anode.*

The capacitance $C_s$ is the space-charge component of grid input capacitance, and is measured as the increase in input capacitance of the valve from cut off to specified working conditions. These measurements, performed at a frequency of 1 Mc/s, need not be discussed further, since a report of very similar measurements has already been published.[7] The mutual conductance was measured under the given working conditions at a low frequency. It is known that neither the capacitance change nor the magnitude of the mutual conductance varies appreciably up to quite high frequencies.

For convenience, the anode shot-noise density (36) was measured at 45 Mc/s. Though this is known to increase considerably at extremely high frequencies, it is not to be expected that there will be appreciable error in assuming that the measured values hold up to 200 Mc/s, at which frequency the grid-noise measurements were performed.

The measured values of mutual conductance $g_m$ and of anode current $I_a$ were also used to determine the equivalent 3/2 power-law transit angle $\omega\tau$ involved in the estimate of grid noise of Section 4. The cathode/grid transit time $\tau$ is given, from 3/2 power-law theory for plane triodes[3], as

$$\tau^4 = 5.34 \times 10^{-41}\, \mu_g\, A^2\, (g_m\, I_a)^{-1}$$

where $\mu_g$ approximates to the value 1 for a fine-

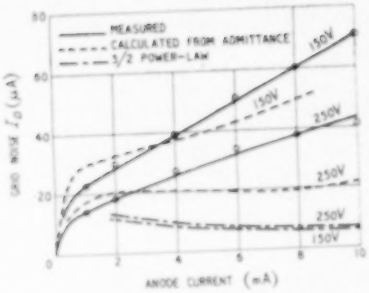



*Fig. 5. Induced grid noise in DET 22, No. 3136, at 200 Mc/s; voltages indicated are those on the anode.*

mesh grid, and $A$ is the electrode area.

We have no accurate method so far of determining the value or behaviour with working conditions of the factor $G^2$ in (35). It is possible, by making various highly arbitrary assumptions and approximations, to deduce from measurements of 'total emission' induced grid noise[8] and an associated admittance change, a qualitative estimate of the behaviour of $G^2$ with anode current and voltage. It would appear from these measurements that $G^2$ is, in fact, scarcely influenced by changing electrode potentials, but this is by no means certain.

## 8. Results

In Figs. 3–6 values of induced grid noise measured on typical high-slope microwave triodes are compared with values predicted from the admittance relation (10) and estimated from the three-halves power-law relation (17). It is assumed that the power spectrum of the fluctuations may be expressed in the form

$$s(\omega) = 2e\, I_D$$

results being given in terms of the pure shot noise in an equivalent temperature-limited current $I_b$.

It is to be seen that for two E1600 valves measured there is a striking similarity between the functional dependence on valve parameters

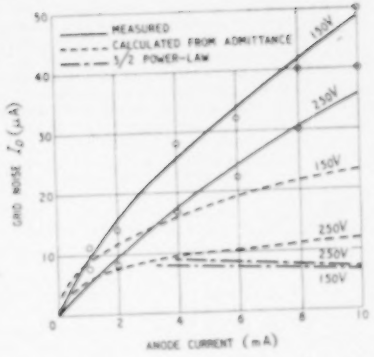



Fig. 6. *Induced grid noise in DET 22. No. 3135, at 200 Mc/s: voltages indicated are those on the anode.*

(anode current and voltage) of the measured and admittance-estimated values of induced grid noise. The 3/2 power-law estimation on the other hand is seen to bear little resemblance to observable fluctuations in any respect.

The DET 22 type of triode, represented in Figs. 5 and 6, has a very much coarser grid than the experimental type E 1600 (grid-pitch to clearance greater) and cannot be expected to follow closely any of the assumptions of internal uniformity which have so far been found necessary in the theory. Considered in this light and bearing in mind the possibilities of considerable experimental error as discussed in Section 6, features of agreement between the predicted and measured curves of Figs. 5 and 6 appear all the more significant.

Reconsideration of Section 6 and, in particular, of Equ. (34) shows that experimental error is qualitatively of the same nature as the discrepancies observable between measured and admittance-predicted grid noise, and may be expected to be large. Thus it is not inconceivable that the predicted results are in fact more reliable as to magnitudes than the measured values themselves.

## 9. Acknowledgments

The work leading to this report was completed by the author as a student of the University of Durham working at the Research Laboratories of The General Electric Company Ltd., Wembley. He has therefore to acknowledge his indebtedness both to the Manager of the Research Laboratories and to the Board of the Faculty of Applied Science, Durham University, for the extension of invaluable facilities.

## REFERENCES

[1] N. R. Campbell, V. J. Francis, E. G. James, *Wireless Engineer*, 1945, Vol. 22, p. 333.
[2] R. L. Bell, *Wireless Engineer*, 1948, Vol. 25, p. 294.
[3] N. R. Campbell, V. J. Francis, E. G. James, *Wireless Engineer*, 1948, Vol. 25, p. 148.
[4] C. J. Bakker, *Physica*, 1941, Vol. 8, p. 23.
[5] D. O. North, *Proc. Inst. Radio Engrs*, 1941, Vol. 29, p. 52.
[6] D. O. North, W. A. Harris, B. J. Thompson, *R.C.A. Review*, 1939, Vol. 4, p. 269.
[7] B. L. Humphreys, E. G. James, *Wireless Engineer*, 1949, Vol. 26, p. 26.
[8] A. van der Ziel, A. Versnel, *Philips Research Reports*, 1948, Vol. 3, No. 1, R.67.

## APPENDIX

*Induced Grid Noise and Noise Factor*

It is interesting to note that the 'distributed' nature of the correlation between induced grid currents and anode shot-noise currents has an effect identical with that of partly *uncorrelated* grid currents (such as the so-called 'total emission grid noise') as far as the operation of the valve as an amplifier is concerned. This may be shown roughly as follows, taking the simple case of the common-cathode triode. Suppose our total cathode-anode shot effect may be represented by

$$i_a(\omega) = \sum_n |i_n(\omega)|^2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

where the $i_n(\omega)$ are elementary shot-effects currents due to the separate electron groups. Then according to the treatment above, the induced noise observable at the grid is of the form

$$i_{gg}(\omega) = \sum_n |\rho_n(\omega) \cdot i_n(\omega)|^2 \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

If there is a current gain $G$ between grid and anode, the net noise spectrum at the anode terminal is

$$i_a = \sum_n |i_n + G\rho_n i_n|^2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

$$= \sum_n |i_n|^2 |1 + (\rho_{nr} + j\rho_{ni})(G_r + jG_i)|^2 \quad \ldots \quad (4)$$

which on expanding and collecting terms becomes

$$i_a = \sum_n |i_n|^2 |1 + 2\rho_{nr} G_r - 2\rho_{ni} G_i + |\rho_n G|^2| \quad (5)$$

By proper choice of $G_r$ and $G_i$ this may clearly be made less than the value

$$\sum_n |i_n|^2 |1 + |\rho_n G|^2| \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

pertaining to a totally uncorrelated grid fluctuation of the above magnitude

$$\sum_n |i_n \rho_n|^2$$

Considering now the hypothetical case where some of the grid currents are *uniquely* correlated with the anode currents ($\rho$ independent of $n$) and there exist other grid currents $\alpha^2 \sum_n |i_n|^2$ uncorrelated with anode currents. We have then an output noise spectrum of the form

$$i_a = |1 + \rho G|^2 \sum_n |i_n|^2| + \alpha^2 |G|^2| \sum |i_n|^2| \quad \ldots \quad (7)$$

which by analogy with (5) can be written

$$i_a = |1 + 2\rho_r G_r - 2\rho_i G_i + |\rho G|^2| \sum |i_n|^2| + \alpha^2 |G|^2| \sum |i_n|^2|$$

The conditions for minimum noise output on varying the real and imaginary parts of the gain $G$ are

$$2\rho_r + 2G_r |\rho|^2 + 2\alpha^2 G_r = 0$$

$$\text{i.e., } G_r = \frac{-\rho_r}{|\rho^2| + \alpha^2} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

$$\text{and} \quad -2\rho_i + 2G_i |\rho^2| + \alpha^2| = 0$$

i.e., $G_r = \frac{\rho_r}{|\rho^2| + x^2}$ .. .. .. .. .. (9)

Substituting in (7), the minimum value of noise output due to the valve is given by

$$e_n = \frac{x^2}{x^2 + |\rho^2|} \sum_s |i_n^2| \quad .. \quad .. \quad .. \quad .. \quad (10)$$

This minimum has the value zero if the *independent* noise source at the grid is zero; i.e., if $x^2 = 0$.

In general, minimizing (5) with respect to $G_r$ and $G_i$, one has

$$\sum_s |i_n^2| \{2\rho_{sr} + 2G_r |\rho_n|^2\} = 0$$

i.e., $G_r = -\frac{\sum \rho_{sr} |i_n^2|}{\sum |\rho_n i_n|^2}$ .. .. .. .. (11)

Similarly, $G_i = -\frac{\sum \rho_{si} |i_n^2|}{\sum |\rho_n i_n|^2}$

Substitution of (11) and (12) in (4) will clearly give a value for the output noise which is less than that given by (6), but this value *will not in general be zero*.

Thus we see that, for the two models to give identical results, the assumption of a simple unique correlation between anode and grid currents implies also the assumption of an independent internal source of noise at the grid. It is clear that noise phenomena which require the invocation of uncorrelated "total emission" components of grid noise are, in fact, more accurately explicable in terms of a continuous amplitude- and phase-distributed, but nevertheless complete, correlation between the currents.

# NEW BOOKS

## Electric and Magnetic Fields

By S. S. Attwood. Pp. 475 + xi, with 276 illustrations. Chapman and Hall, Ltd., 37 Essex St., Strand, London, W.C.2. Price, 44s.

The author is a professor of electrical engineering at the University of Michigan and this is the third edition of the book, which was first published in 1932. In the preface the author says that "in spite of the time devoted during the first two years to the sciences which are basic for electrical engineering, the student needs further training in the development of the fundamental concepts, formulas, terminology, and units used in electric and magnetic field study." This book is intended to provide this training. It is divided into four parts, the first dealing with electric fields, the second with magnetic fields of constant permeability, the third with ferromagnetic fields, and the fourth of only 34 pages with combined electric and magnetic fields; that is, with electromagnetic waves.

Although some parts of the magnetic section have been rewritten to emphasize currents rather than poles, "the magnetic pole concept has not been abandoned entirely since its value as a pedagogical tool is too well established." Considerable attention has been given to field mapping and many diagrams are given of electric and magnetic fields showing the lines of flux and the equipotential surfaces. "A by no means unimportant benefit that should accrue to the student from the presence of many carefully developed field drawings is the stimulation of his imaginative sense." The rationalized m.k.s. system of units has been adopted but tables are given showing the relations between the units in the various systems.

There is a very strange arrangement of material at the beginning of the book. The first eight pages deal with units and dimensions, and if one then jumps to page 23, one proceeds in a normal manner with "elementary facts and definitions," but pages 9 to 22 are devoted to such things as the electronic energy levels in the hydrogen atom, the Lyman, Balmer and Paschen series, and Maxwellian velocity distribution curves. It is difficult to see why this strange material was inserted at this point between the definitions of the international ohm and ampere on p. 8 and Coulomb's law on p. 24. There are other things about the book that are unsatisfactory, and it is surprising to find them in a third edition. In the diagram showing the dependence of dielectric constant and losses on frequency, the dielectric constant is shown going down to 0.5 at certain frequencies, and a curve is stated to give the loss per cycle although the ordinates are labelled power; even when the loss per cycle is shown as constant over a wide range of frequencies, the total loss is also shown as constant and not increasing with the frequency. All this should be carefully explained or omitted.

On looking into the calculation of the force between two parallel conductors and their inductance, one finds the same unsatisfactory treatment. The wire radius is assumed to be small, and the formula for the force is stated to be for $R \ll D$, whereas it applies for any size of wire and the assumption is quite unnecessary and misleading, as is obvious from the fact that the flux density in which one wire is situated, and therefore the force on it, is independent of the size of the other wire. The calculation of the inductance is quite amusing; having calculated the inductance of one wire due to its external flux, the author says "When the flux inside the wire is considered it is difficult to determine the number of flux linkages, so we resort to the energy definition." Here again the mistake is made of assuming that the formula is only correct when $R \ll D$, whereas, if the current density is uniform, the formula is quite correct whatever the size of the conductors, as can be easily shown. It is simpler to assume that the conductors have the same permeability as the surrounding space; any marked departure from this is so rare that it need not be considered in such a book.

The book concludes with a very interesting 8-page historical outline of electricity and magnetism from the earliest times down to the present day.

G. W. O. H.

## The Principles of Television Reception

By A. W. Keen, M.I.R.E., A.M.Brit.I.R.E. Pp. 319 + xv. Sir Isaac Pitman & Sons Ltd., Parker St., Kingsway, London, W.C.2. Price 30s.

An unusual feature of this book is the extent to which the American television standards are treated. Throughout, there is a parallel treatment of circuit detail from both the British and the American viewpoints. While this is helpful in showing very clearly how the different transmitting standards affect receiver design, it does make it rather more difficult for those interested primarily in only one system to obtain a comprehensive picture of that system.

The book opens with a table of British and American technical terms. This is a good point, for it emphasizes the differences of language which are liable to be confusing—especially the American use of "frame" to mean picture. The first chapter describes in simple language the basic principles of television including the scanning-

process and the British and American waveform standards.

The next chapter covers a wide field, for not only does it deal with "waveforms" and the circuit response to them but the basic r.f. amplifier circuits, including the earthed-grid stage, are also treated. As one might expect there is no analysis of the circuits but merely an elementary description of how they work. The transient response section, however, is quite well done and gives a very good account of the effect of simple circuits upon pulses of various kinds.

Chapter III covers the c.r. tube with its focusing and deflecting arrangements and, again, a good general picture of the subject is given without going into it very deeply. E.h.t. supplies are also discussed. Chapter IV is devoted to time bases and covers, not only the saw-tooth and pulse voltage generators, but current amplifiers for magnetic deflection and linearizing circuits. A chapter on sync separation follows.

A single chapter covers the whole of vision-signal amplification in only 40 pages. As v.f., i.f. and r.f. amplifiers, detectors, gain control, sound-channel rejection, frequency-changers and noise are covered, it is plain that the treatment can only be superficial.

The sound channel has a chapter to itself, but is not much more detailed since, in addition to the British a.m. and the American f.m. systems, methods using pulse-width modulation are also described.

There is a lengthy and well-illustrated chapter on the complete receiver in which circuit diagrams are given of several British and American sets. Aerials and feeders have a short chapter, receiver test equipment another and one on colour television ends the book proper. There is a short appendix containing a number of useful mathematical expressions. Each chapter contains a useful, if condensed, bibliography.

In his preface, the author states that the book has grown out of material originally produced as a short correspondence course for engineers of a receiver manufacturer's servicing organization. It is certainly not a servicing book and is not intended to be. It is meant to cover principles only and it does so. Whether or not it does so adequately is another matter and a debatable one, since it depends on how much one expects a service technician to know. In the reviewer's opinion it does no more than lay a foundation for further study.

Mathematics are almost non-existent in the body of the book and even in the Appendix are quite simple. The book can be recommended for those who have little or no knowledge of television but who have a good background of ordinary wireless theory and practice. It will enable them to obtain a good general, but hardly a detailed, knowledge of television. The book is not entirely free from errors, but where they are not obvious they are unlikely seriously to hamper the student.

W. T. C.

**Quirlende elektrische Felder**

By Fritz Emde. Pp. 118 + vii, with 41 illustrations. Friedr. Vieweg und Sohn, Braunschweig, Germany.

Dr. Emde, who was formerly Professor of Electrical Engineering at Stuttgart, is known to a wide circle outside the field of electrical engineering through his Tables of Mathematical Functions. This present small volume on electric fields was handed over to the publishers in 1943 but has only just been published. It is confined to whirling or curl fields, that is, to the relatively small fields associated with magnetic fields, and does not deal with the powerful electric fields occurring in high voltage work. The author states that his main object is to give the reader a clear conception of the principles underlying the fundamental laws of Maxwell, Heaviside, etc.

The treatment is very full and clear and illustrated by simple numerical examples. On p. 48 a field is resolved into its two components, one a curl field and the other a curl-free field; this is done very simply for the case of a packet of iron laminations with numerical examples. Several well-known paradoxes, such as the revolving bar magnet and Hering's spring tongs, are examined on p. 72. Several pages are devoted to the movement of dielectrics in a magnetic field, and Wilson's classical experiments are described. The final twenty pages are more mathematical, dealing with divergence, rotors, Spreizung, Drillung and Affinors, but in the introduction Dr. Emde says that he trusts that those readers who put the book down without working through it to the very end will have gained something from it. The author decided to omit electromagnetic waves as they are treated so fully in a number of books. This is a book that can be unreservedly recommended to anyone with a knowledge of German.

G. W. O. H.

**Ein Ultrakurzwellen-Telefoniesystem hoher Kanalzahl mit Frequenzweiche**

By G. C. Fontanellaz. Pp. 74, with 41 illustrations. Verlag Leeman, Zürich. Price 10 francs (Swiss).

This is the author's doctorate thesis at the Zurich Hochschule, on work carried out under the guidance of Dr. Tank.

The first 30 pages are devoted to a discussion of the general problem of multiple-channel telephony, with a very large number of channels. The remainder of the book contains a description of a proposed system which was constructed and tested in the laboratory. Photographs of the actual apparatus are given. The system contains 6 u.h.f. bands, each with 12 speech channels; the aerial is of the highly-directive narrow-beam type. The frequencies discussed are between 200 and 3000 Mc/s. Various methods of producing and modulating the pulses are discussed. It should be noted, however, that this is the fourth monograph published by the Hochschule during 1949 dealing with closely interlinked subjects, or branches of the same subject, and anyone interested in the subject would be well advised to obtain the four monographs. The others were "Ultrakurzwellen Frequenzweiche," by Staub (7.50 f), "Breitband Richtstrahl Antennen," by Peter (8 f), and "Impulsmodulation," by Bachmann (9 f).

G. W. O. H.

**Stetige Regelvorgänge** (Servo-mechanisms)

By Winfried Oppelt. Pp. 144, with 42 illustrations. Wissenschaftliche Verlagsanstalt K.G., Hannover. Price DM.7.80.

In March 1948, p. 88, we reviewed a book by the same author entitled "Grundgesetze der Regelung" and stated that it contained the basic mathematical equipment, the practical application of which would be dealt with in another volume. This is the volume referred to. It deals with every possible type of servo-mechanism, mechanical, electrical and thermal, with diagrams of the mechanism, and numerous graphs illustrating their operation, stability, etc. The book concludes with a very full bibliography of the subject.

The book was completed, and ready for publication, in 1945; the author states in the preface that subsequent developments are of course, not dealt with in the text, but the bibliography has been brought up to date. The new volume is not an austerity production as was the first; the production, and especially the illustrations, are excellent and the two volumes form a valuable introduction to a subject of growing importance. We commend them to anyone with even a small knowledge of German; so much of the material is graphic.

G. W. O. H.

# SIGNAL-TO-THERMAL NOISE RATIO

## *Comparison of F. M. and A. M. Receivers*

By M. V. Callendar, M.A., A.M.I.E.E.

**SUMMARY.**—Curves are given for output signal/noise ratio for thermal noise with an f.m. receiver, including conditions where the signal is equal to or less than the noise, as well as the well-known case where the signal greatly exceeds the noise. Corresponding curves are also given for an a.m. receiver, and a comparison is drawn between the results for f.m. and a.m., bringing out especially the effect of the ratio of pre-detector to post-detector bandwidth in each case.

It is concluded that, apart from its well-known advantages for high-quality broadcasting, f.m. can give results at least equal to a.m. in respect to signal/noise ratio for communication work provided that the pre-detector bandpass is no wider than necessary to accept the full deviation, and that the limiter is fully efficient: the deviation ratio for such work should not exceed about four. However, these conditions are not always easy to comply with in practice (for example, in cases where a wide bandwidth is necessary from frequency-drift considerations) and thus an a.m. receiver will, in some instances, be found to give a slightly better range of intelligible reception than its f.m. counterpart, particularly if any slight mistuning be present.

### 1. Introduction

THE question of signal/noise ratio in f.m. receivers has been discussed by many writers since Armstrong's original paper[1]. In that paper, the improvement in signal/noise ratio for an f.m. receiver as compared with an a.m. receiver was evaluated for the case where the signal at the detector is considerably greater than the noise; this analysis is all that is necessary when considering good quality broadcast reception, but sheds little light on the important question of the relative merits of f.m. and a.m. in respect of maximum range of intelligible communication. In a recent paper[2] the present author evaluated formulae for signal/noise ratio in a.m. receivers and examined its variation with bandwidth. Since that paper was written a paper[3] by Stumpers on noise in f.m. receivers has been published; on the basis of his results we can calculate signal/noise curves for f.m. down to below the noise level, and thus compare f.m. and a.m. receivers at any level, and for any bandwidth.

As regards the requirements in audio signal/noise ratio for the various types of service, the following table may be taken as a fair indication:

| | |
|---|---|
| Telegraphy | 0 db |
| Bare minimum intelligibility (telephony) | 10 db |
| Comfortable intelligibility (telephony) | 20 db |
| Minimum acceptable for broadcast listening | 30 db |
| Good quality broadcast reception | 40 db |

### 2. Theory

The results of Stumpers' excellent mathematical paper[3] are expressed in a series of curves for a.f. noise output against noise-to-signal ratio at the input to the detector, for various value of the ratio $\Delta f/f_a$ where $2\Delta f$ is the pre-detector bandwidth and $f_a$ is the audio bandwidth. The detector is here assumed perfect; i.e., it responds

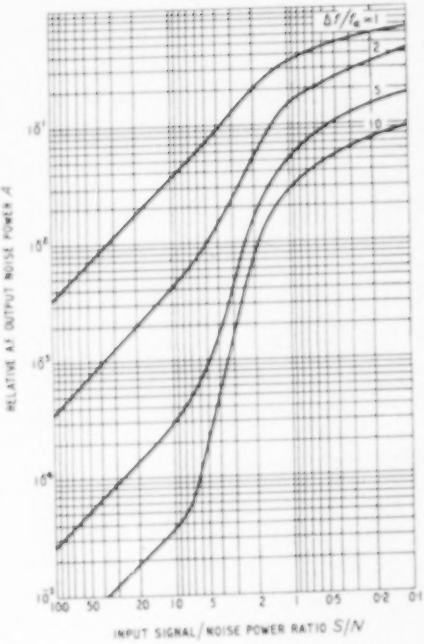



Fig. 1. *Variation of output noise from an ideal f.m. detector with input signal/noise ratio. For a rectangular pre-detector filter of bandwidth $2\Delta f$ and audio band $f_a$. From Stumpers[3] (Figs. 2 and 3), who gives noise output* $= A\Delta f^2/3$.

MS accepted by the Editor, July 1949

only to changes in frequency and not to changes in amplitude. Curves for the case of a rectangular pre-detector bandpass characteristic are reproduced here as Fig. 1, and clearly form a basis from which signal/noise ratio can be obtained: in their existing form, however, they are apt to give a misleading impression of the relation of signal/noise ratio to bandwidth since the 'noise-to-signal ratio' used as abscissa is itself a function of bandwidth (see below).

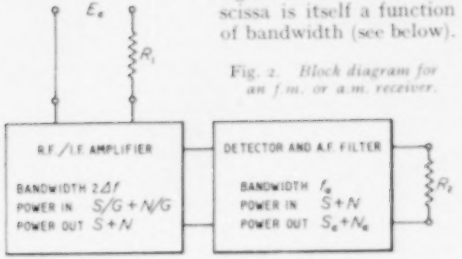



Fig. 2. *Block diagram for an f.m. or a.m. receiver.*

Stumpers also gives a formula for the suppression (due to limiter action) of the signal modulation by the noise when the noise-to-signal ratio is large; the amplitude of the signal modulation is reduced in the ratio $B = (1 - e^{-S/N})$ where $N$ is noise and $S$ is signal, both in terms of power at the input to the detector.

Assuming that the frequency detector is 'perfect' (i.e., that it includes a perfect limiter as well as the usual frequency to amplitude converter and amplitude detector) we may define a conversion factor $C$ such that the detector gives an output of $C$ watts per cycle deviation at the input. The assumption that $C$ is invariable under all conditions (of input amplitudes, etc.) is made by Stumpers; this assumption is evidently not true if the input to the detector diodes is too small, but this is perhaps unlikely in practice (since the detector follows the limiter). However, it is necessary to take into consideration the suppression of modulation by noise which must occur in the diodes which 'convert' the i.f. signal (here amplitude modulated) from the discriminator into the required audio signal. This correction factor $D$ is small and only appreciable for very low signal/noise ratios ($S/N <$ say 6 db) where we may assume the output noise spectrum roughly uniform (Stumpers' Fig. 6): thus we can put $D = 1/(1+N'/2S')$ approx. (vide formula for a.m. below) where $N'/S'$ is the output noise/signal ratio for $\Delta f = f_a$ as calculated direct from Stumpers.

Then signal a.f. output power =

$$S_a = m^2\delta f^2 \times C \times D \times B^2$$

where $\delta f$ is peak deviation, $m$ is fractional modulation, and $B$ and $D$ are modulation-amplitude suppression-ratio factors for the limiter and the detector respectively. And noise a.f. output power $= N_a = A \times \Delta f^2/2$ multiplied by $2C$, since Stumpers assumes a d.c. output of 1 V across 1 Ω for 1-c/s deviation. Here $A$ is the ordinate in Fig. 1, and $2\Delta f$ is pre-detector power bandwidth. Thus the a.f. signal/noise ratio =

$$\frac{S_a}{N_a} = \frac{m^2\delta f^2}{\Delta f^2} \cdot \frac{B^2 D}{A}$$

If, as in Fig. 2, a signal e.m.f. ($E_a$) from a source of resistance $R_1$ is applied to a receiver having a total pre-detector power gain $G$, the signal power at the input to the detector will be $G \times E_a^2/4R_1$, assuming matching of the receiver to the source. The corresponding noise power will be $G \times FKT \cdot 2\Delta f$, where $F$ is the receiver 'noise factor' and $KT$ has the usual significance (noise power available from a resistance per cycle bandwidth).

Thus the signal/noise power ratio at the input to the detector $= \dfrac{S}{N} = \dfrac{E_a^2}{8FKT\Delta fR_1}$

For convenience, we will here define a parameter $S_r = \dfrac{S}{N}\dfrac{\Delta f}{f_a}$ and call this the "relative signal input." Then we have $S_r = \dfrac{E_a^2}{8FKTR_1 f_a}$ and this represents the signal relative to the

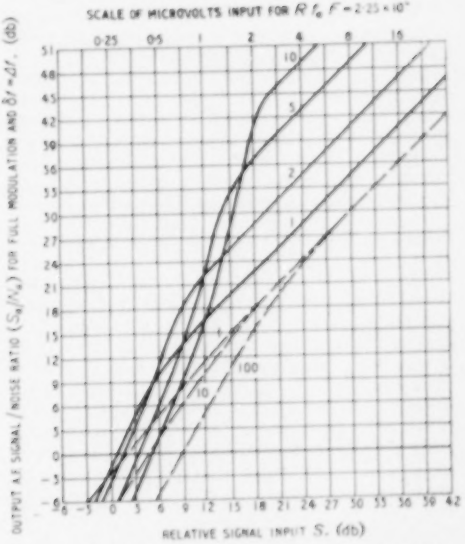



Fig. 3. *Signal/noise ratios for an ideal f.m. receiver compared with those for an a.m. receiver. Number marked on curves is ratio of half pre-detector bandwidth to post-detector bandwidth ($\Delta f/f_a$). Full lines for f.m., dashed lines for a.m.*

noise in a bandwidth $2f_a$. We can now plot curves (Fig. 3) for a.f. signal/noise ratio against $S_r$, which is independent of the pre-detector bandwidth: for simplicity, we take the case where $m = 1$ and $\Delta f = \delta f$: the actual signal/noise ratio for any lower deviation will be proportionally less.

For comparison, we have also plotted the signal/noise ratio for an a.m. receiver from the relation:

$$\frac{S_a}{N_a} = \frac{m^2S}{N} \cdot \frac{\Delta f}{f_a} \cdot \frac{1}{(1+N/2S)} = \frac{m^2S_r}{(1+N/2S)}$$

This is taken from Section 2.5 of the paper[2] by the present writer cited above; it refers to a parabolic detector but holds also for a linear detector to within, say, $\pm$ 1 db at all inputs. In the case of a linear detector, the factor $1/(1 + N/2S)$ allows for the suppression of modulation by noise when the signal/noise ratio is low: with a parabolic detector, the same factor merely expresses the quadratic relation between input and output.

The curves as plotted give all the information necessary to compare theoretical signal/noise ratios for a.m. and f.m. for any ratio of pre-detector to post-detector bandwidtn; such comparisons will be unaffected by the a.f. bandwidth, the noise factor, or the signal-source impedance. However, if we require the numerical signal/noise ratio for any particular case, it can readily be re-plotted in terms of microvolts input ($E_a$) from the equation given above relating $S_r$ to $E_a$ for any given values of $F$, $R_1$ and $f_a$. A scale of microvolts for the particular case of a typical v.h.f. receiver having $R_1 = 75\,\Omega$ $F = 10$ and $f_a = 3$ kc/s, has been added to the figure by way of illustration.

When the signal is much larger than the noise ($S/N > 10$) the curves conform to the simple formula $\frac{S_a}{N_a} = 3 \times \frac{S}{N} \times \frac{\delta f^2}{f_a^2}$ which agrees with Armstrong's original formula, for signal/noise ratio. Fig. 4, derived directly from Fig. 3, gives the effect of increasing the pre-detector bandwidth without altering the deviation, and agrees with the above simple formula in showing that when $S/N <$ about 10, an increase in pre-detector bandwidth does not affect the a.f. signal/noise ratio for a given deviation ratio.

It should be noted that the curves of Fig. 1 assume a rectangular selectivity curve for the pre-detector circuits: Stumpers also gives figures for a curve of Gaussian form, but the results from this are not substantially different and do not, in the present writer's view, approximate any more nearly to practical conditions. In practical cases, where the curve is not rectangular, $\Delta f$ must be taken as the 'power bandwidth' defined by $2\Delta f = \frac{1}{S_{max}} \cdot \int_{-\infty}^{+\infty} S.\, df$

A further limitation on accuracy is that the noise output is not strictly independent of deviation of the carrier: (see Stumpers[3] p. 1090). It is felt that it is justifiable to neglect this second-order effect in practice, since it will only occur at large deviations where the noise will be usually masked by the modulation. This effect should not be confused with the large effects of deviation or detuning when the limiter is not in operation and the balance of the discriminator is relied upon for noise reduction. However, it does mean that signal/noise ratio deteriorates appreciably with mistuning even with a perfect limiter in action.

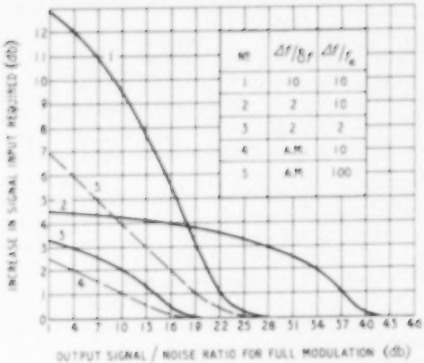



*Fig. 4. Increase in signal input required to restore output signal/noise ratio if pre-detector bandwidth is made wider than ideally necessary (i.e., if $\Delta f > f$ for f.m. or if $\Delta f > f_a$ for a.m.). Full lines for f.m., dashed lines for a.m.*

Curves are also given by Stumpers for the case where *pre-emphasis* at the transmitter enables a de-emphasis filter to be inserted at the receiver. For the case of a high fidelity system where $f_a = 15$ kc/s and the filter time constant is 75 microseconds (3 db cut at 2 kc/s), the output signal/noise ratio is improved by 12 db approx. when $S/N$ is $> 10$. For lower values of $S/N$, the output noise spectrum progressive changes from one increasing linearly with frequency to one roughly uniform over the audio band (and thus similar to that obtained on a.m.): the improvement from pre-emphasis thus becomes smaller and eventually falls to about 6 db for $S/N = 1$ on f.m., where it is equal to that obtained for a.m. For a communication system having $f_a = 3$ kc/s, since the maximum permissible pre-emphasis is about 3 db at 1.5 kc/s (time constant 100 microseconds) it is clear that the gains in signal/noise ratio will be much less and for an output signal/noise ratio of 10 to 20 db (i.e., for $S/N$ values

around 1 to 2) the gain from pre-emphasis will be small (about 2 db) and not materially different for a.m. and for f.m.

### 3. Practical Conclusions

It is important to emphasize at the outset that, while there is no reason why the curves of Fig. 3, for a.m. should not hold under practical conditions, the curves for f.m. represent an *ideal receiver*, and results from actual receivers will be worse for two reasons :—

(a) It is assumed that the peak deviation ($\delta f$) can be made equal to half the power bandwidth. Even for a good 'square' bandpass, this would entail loss of linearity for high modulations and, in practice, a $\Delta f$ of $1.3\delta f$ to $2\delta f$ must be allowed for. From Fig. 4, an increase of 1 to 4 db in signal input will be required to restore the a.f. signal/noise ratio when the latter is of the order of 10–20 db. In certain cases, notably where a small deviation is used at v.h.f., the i.f. bandwidth will have to be made much greater than the deviation in order to cover possible frequency drift : here the signal/noise ratio will be noticeably worsened.

(b) The detector will not necessarily be completely insensitive to the amplitude modulation energy in the noise. The correction required here will be less than 1 db with really good limiting and accurate tuning, but may be quite large for small signals in receivers of inadequate sensitivity: in such cases, the signal/noise ratio will be sharply dependent upon the exact centering of the carrier and the symmetry of the tuned circuits, and the noise will increase notably with modulation.

An increase of 3 db in signal at small inputs, falling to 0 db at larger inputs (see Fig. 4) is considered adequate to cover the above factors for a really *efficient practical f.m.* receiver : our conclusions for such a receiver are expressed in the table.

TABLE

R.F. INPUTS required for a given output signal/noise ratio (expressed relative to an a.m. receiver having $\Delta f = f_a$)

| | $\Delta f/f_a$ | Required A.F. Signal/Noise Ratio | | | |
|---|---|---|---|---|---|
| | | 0 db | 10 db | 20 db | 40 db |
| Amplitude Modulation | 1 | 0 db | 0 db | 0 db | 0 db |
| | 10 | +3 | +1 | 0 | 0 |
| | 100 | +7 | +4 | +1 | +0 |
| Frequency Modulation | 1 | +2 | −2 | −4 | −4 |
| | 2 | +3 | −2 | −9 | −10 |
| | 5 | +5 | −1 | −8 | −18 |
| | 10 | +6 | +2 | −6 | −22 |

In this table, comparisons are of course only to be made in the vertical columns : the deviation ratio may be taken as three-quarters of the band ratio quoted in the table.

As regards, then, a communications receiver for telephony, we conclude :—

(i) The f.m. receiver should be equal to, or slightly better than, the a.m. receiver as regards range of intelligible reception, *provided* that it has (a) a limiter fully operative at the smallest inputs, (b) an i.f. curve shape not far from rectangular and an i.f. bandwidth not exceeding 3 times the deviation, (c) a suitable deviation ratio (not exceeding 4). A small improvement (some 2 db), may be obtained for both a.m. and f.m. by pre-emphasis. Accurate tuning is assumed.

(ii) If, however, these conditions are not satisfied, the a.m. receiver will tend to have an advantage. In particular, the effective sensitivity of an a.m. receiver is but little affected in most practical cases by the pre-detector bandwidth (see previous paper[2] for examples) but, in the case of f.m. communication receivers, it is essential that the pre-detector bandwith be kept down to the minimum compatible with the requirements of the signal deviation. As an example, an increase in bandwidth from $\Delta f = 5f_a$ to $\Delta f = 10f_a$ could lower the a.f. signal/noise ratio from 18 db to 6 db (for $m\delta f = 4f_a$ and $S_r = 10$ db). For high quality broadcast reception, however, the pre-detector bandwidth ceases to be important, as indicated by Fig. 4.

As an extreme case where a.m. will give superior signal/noise ratio and range, we may take a miniature portable receiver on, say, 160 Mc/s, where the normal $\pm$ 0.01% overall frequency stability on both receiver and transmitter imposes a bandwidth of at least 80 kc/s for − 3 db and gain is insufficient (due to space and weight limitations) to permit of efficient limiting below, say, 10 $\mu$V. Accurate centering of the receiver discriminator on the transmitter is likewise impractical and it is clear enough that the input for 10 db signal/noise ratio for f.m. will be appreciably greater than that for a.m., whether wide or narrow deviation is used. At the other extreme, of course, we have the usual high quality broadcast receiver, where the range for 40 db audio signal/noise ratio will be nearly three times better with f.m. than with a.m.

In the only practical example in which the writer has made actual comparative measurements (a battery-operated Services f.m. set), it was found that an input of 1.2 $\mu$V was the minimum for intelligible signals with f.m. and gave an a.f. signal/noise ratio of 0 db for 30% modulation : the same figure for minimum input was obtained on changing the detector to operate on a.m. In this case, the limiter was not fully operative at the inputs in question and the i.f. curve was that

obtained from four tuned circuits in cascade: $\Delta f$ was about 15 kc/s and $\delta f$ was about 10 kc/s.

If the receiver was slightly detuned, results were of course worse on f.m. than on a.m.

Since the above was written, a paper entitled "Theoretical Signal-to-Noise Ratios in F.M. Receivers" by D. Middleton, has appeared in the *Journal of Applied Physics* for April 1949. This paper treats the subject exhaustively from a mathematical point of view; the practical aspects are touched on somewhat briefly, and the conclusions are slightly different from those in the present paper, owing to different assumptions (mainly a Gaussian shape of response characteristic in place of a rectangular shape).


REFERENCES

1 Armstrong. "Frequency Modulation," *Proc. Inst. Radio Engrs.*, May 1936.

2 Callendar. "Thermal Noise in A.M. Receivers," *Wireless Engineer*, Dec. 1948.

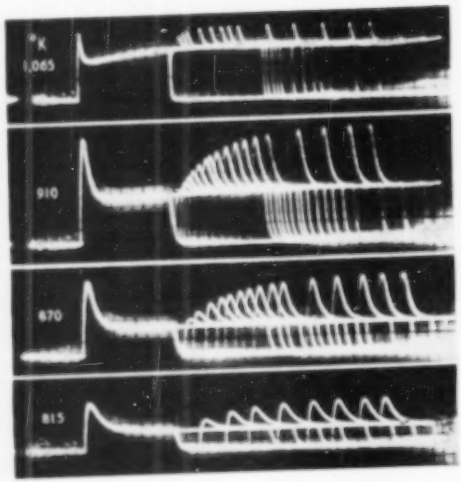
3 Stumpers. "Theory of F.M. Noise," *Proc. Inst. Radio Engrs*, Sept. 1948.


# CORRESPONDENCE

*Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.*

## Change of Mutual Conductance with Frequency

SIR,—The October 1949 *Wireless Engineer* published a paper[1] on this subject which was based on the operating characteristics of pentode-type valves. New valves exhibited little frequency dependence of the mutual conductance, $g_m$, but valves operated over 1,000 hours showed marked reduction in $g_m$ with decreasing frequency between $10^7$ and $10^8$ c/s. This frequency dependence of gain was explained as arising from a parallel resistance-capacitance impedance in the cathode circuit causing negative feedback. This impedance was associated with a change in the physical nature of the oxide cathode interface. The hypothesis was offered that the electrical contact between the oxide coating and the base metal deteriorated during the life of the valve. When this occurred only a few points of contact bridged the thin vacuum gap separating the base metal and coating, thus forming effectively a parallel RC impedance to the normal flow of electron current from the base metal into the coating.




*Anode current pulses of 7AD7 pentode with 15-μsec grid pulse driving the grid from — 20V to zero with an anode supply of 400 V.*

A study[2] has been made of the decay of anode current passed by a pentode when a square voltage pulse is applied to the grid. These results were also interpreted in terms of a parallel RC impedance at the oxide cathode interface but the physical model offered to explain the impedance used the development of a homogeneous interface compound, sandwiched between the coating and base metal, having a low conductivity. It seems likely that this rather than the inhomogeneous, poor contact hypothesis is the correct explanation of the change of $g_m$ observed by Raudorf. The figure shows the anode current pulses passed by a type 7AD7 pentode upon the application of a 15-μsec grid pulse sufficient to drive the control grid from cut-off to zero bias. The decay is readily apparent and is more pronounced at lower temperatures. Following the initial pulse a second pulse is applied after a varying time delay. When the time delay is short the initial current is less than that for the first pulse. A number of these for different delays are superimposed to form the figure. The recovery of anode current to the initial value obtainable at the beginning of the first pulse may be seen. The decay constant (time for the plate current to decrease to $1/e$ of its initial value) and the recovery constant, similarly defined, are seen to be different and to be temperature dependent. Using the equations for degenerative feedback amplifiers the two-time constants are:

$$\text{Decay constant,} \quad D = RC(1 + g_m R) \quad (1)$$

$$\text{Recovery constant} \quad D' = RC \quad (2)$$

where $R$ and $C$ are the resistance and capacitance of the interface. From an experimental evaluation of $D$, $D'$ and $g_m$ values of $R$ and $C$ may be obtained. Tabulated below are the results of this study which seem pertinent to the association of the impedance with an interface compound.

1. Cathodes of valves made using a pure nickel base metal do not develop the decay.
2. Cathodes of valves made using a nickel base metal containing 0.1 to 0.2 per cent silicon do develop the decay.
3. X-ray diffraction patterns of cathodes made on silicon nickel show an interface of $Ba_2SiO_4$ approximately $10^{-4}$ cm thick. New cathodes have no decay and no interface compound.
4. The measured interface resistance and thickness are consistent with values[3] of the specific electrical conductivity of $Ba_2SiO_4$. The temperature depend-

ence of resistance is that to be expected of a semiconductor.

5. The measured interface capacitance is independent of temperature and is consistent with the measured interface thickness and an assumed dielectric constant of 10.

Similarities between our observed decay phenomena and the reported frequency dependence of $g_m$ are as follows:

1. Both effects develop during the same life period of the valve.
2. Decay constants of about 1 $\mu$sec are consistent with the frequency dependence between $10^7$ and $10^8$ c/s.
3. Values of $R$ and $C$ obtained using the two methods of analysis are essentially the same, $C \approx 2 \times 10^{-4}$ pF and $R \approx 100\ \Omega$.
4. Cathode sparking has been associated with the development of both interface models, see Ref. 3.

Because of these similarities it is suggested that the two phenomena arise from the same physical model. Since the base metal composition used is not specified by Raudorf it cannot be assumed that the interface compound in his valves is $Ba_2SiO_4$; however, many reducing impurities found in commercial cathode cores are capable of forming other interface compounds.[4] In addition to adequately accounting for the frequency dependence of $g_m$ our model based on a growth in interface thickness with life offers an explanation for the reduction in capacitance with increasing resistance as reported by Raudorf. This was not explained by the poor contact hypothesis which accounts only for the resistance increasing with cathode life.

From the standpoint of practical cathode operation it is of interest to note that the decay phenomena develops more readily[5] in valves operated under cut-off conditions than under class A operation. This we attribute to the development of unactivated $Ba_2SiO_4$ in the former case since the interface is formed in both types of operation. Previous study[3] showed that this interface compound could be activated by electron current flow. A detailed account of these and other experiments is being prepared for publication elsewhere.

Department of Physics,
University of Missouri.

A. EISENSTEIN.

[1] W. Raudorf, *Wireless Engineer*, October 1949.
[2] A. Eisenstein, paper presented at American Physical Society meeting, November 25, 1949.
[3] A. Eisenstein, *J. appl. Phys.* Vol. 20, No. 776, 1949.
[4] D. A. Wright, *Proc. Phys. Soc.* B, Vol. 62, p. 188, 1949.
[5] Results obtained at the M.I.T. Servomechanisms Laboratory. The writer is indebted to E. S. Rich of that laboratory for supplying valves used in this study.

## A Problem — An Answer

In the February issue we published an enquiry from a reader. Mr. A. T. Starr, of Standard Telecommunication Laboratories Ltd., gives the following answer:—

"Regarding the problem of finding the monotonic function with the fastest rate of rise, the solution is very old and is contained in almost any book on Analysis or Operational Calculus. It is well known that

$$\frac{1}{2\pi j}\int_{c-j\infty}^{c+j\infty} e^{\lambda t}\,d\lambda/\lambda = 0,\ t<0,$$
$$= 1,\ t>0.$$

To prove this it is only necessary to complete the contour by a large semi-circle to the right for $t<0$, when the integral is zero: when $t>0$ the semi-circle must be to the left and then there is a pole at $\lambda = 0$ at which the residue is unity. Thus if $f(\lambda) = 1/\lambda$, the time function rises instantaneously from zero to unity at $t = 0$: it is thus monotonic and has an infinite rate of rise.

If I may guess what is in the reader's mind, I should say that he wants the solution of an amplifier with a monotonic response and fastest rate of rise. This problem cannot be said to be solved, but it is possible that the work of W. E. Thomson, *Wireless Engineer*, January 1947, has approached, if not actually obtained, the solution." [Ed.]

## STANDARD FREQUENCY TRANSMISSIONS

From 1st February 1950 standard frequencies are being transmitted from Rugby on 60 kc/s, 5Mc/s and 10 Mc/s with powers of 10 kW. The times of transmission are: 60 kc/s, 1029–1045; 5 Mc/s, 0544–0615; 10 Mc/s, 0629–0700; all times being G.M.T. The accuracy is within two parts in $10^8$ and is monitored by the National Physical Laboratory to which all enquiries and comments should be addressed at Teddington, Middlesex.

Each transmission is modulated with the following 15-minute cycle:—

Modulation, Slow morse call sign (MSF) followed by speech, at 59-00, 14-15, 29-30, 44-45, minutes past the hour.
Modulation, 1,000-c/s tone, at 00-60, 15-20, 30-35, 45-50 minutes past the hour.
No modulation, at 05-14, 20-29, 35-44, 50-59 minutes past the hour.

## INSTITUTION OF ELECTRICAL ENGINEERS

The following meetings will be held at the I.E.E., Victoria Embankment, Savoy Place, London, W.C.2, commencing at 5.30:—

March 15th, "The Structure, Electrical Properties and Applications of the Barium-Titanate Class of Ceramic Materials," Prof. Willis Jackson.

March 27th, "The Operation and Maintenance of Television Outside-Broadcast Equipment," T. H. Bridgewater.

March 30th, Symposium of Papers on the M.K.S. System of Units.

April 3rd, Discussion meeting on "The Place of the Model in Teaching," opened by D. G. Sandeman.

April 12th, "A Review of some Television Pick-up Tubes," J. D. McGee, and "The Design of a Television Camera Channel for use with the C.P.S. Emitron," E. L. C. White and M. G. Harker.

## BRITISH INSTITUTION OF RADIO ENGINEERS

During March the following meetings will be held:—

**London Section.** March 23rd, at 6.30, at the London School of Hygiene & Tropical Medicine, Keppel St., W.C.1: "High Performance Television Monitors," by J. E. Jacobs.

**West Midlands Section.** March 22nd, at 7.0, at the Wolverhampton & Staffordshire Technical College, Wulfruna St., Wolverhampton: "Radio Interference with Broadcast Reception," by A. A. Devey.

**North-Eastern Section.** March 15th, at 6.0, at Neville Hall, Westgate Rd., Newcastle-on-Tyne: "Electrical Breakdown of Gases at Ultra High Frequencies," by Dr. Prowse.

**South Midlands Section.** March 30th, at 7.0, at the Coventry Technical College, Room A5. "Electronics and the Brain," by H. W. Shipton.

# WIRELESS PATENTS

## *A Summary of Recently Accepted Specifications*

*The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 2/- each.*

### ACOUSTICS AND AUDIO-FREQUENCY CIRCUITS AND APPARATUS

627 639.—Moving-coil loudspeaker in which the output transformer acts as the field-producing magnet; this is stated to give a volume expansion effect.

*E. K. Cole Ltd. and E. L. Hutchings. Application date April 3rd, 1947.*

### AERIALS AND AERIAL SYSTEMS

627 689.—Aerial for ships for use at u.h.f. employing conductive cones designed to give omni-directional radiation.

*Marconi's Wireless Telegraph Co. Ltd. (assignees of J. H. Atwood). Convention date (U.S.A.) May 4th, 1942.*

627 091.—Tunable waveguide in which the tuning cup is associated with a loading element adjustable relatively to the cup.

*Hazeltine Corporation (assignees of J. A. Rado). Convention date (U.S.A.) March 30th, 1946.*

627 092.—Waveguide in which the tuning cup is supported by a dielectric ring bearing against a conducting surface of the guide.

*Hazeltine Corporation (assignees of J. A. Rado). Convention date (U.S.A.) March 27th, 1946.*

### RECEIVING CIRCUITS AND APPARATUS

*(See also under Television)*

627 238.—Stability is ensured in a high gain i.f. amplifier by a frequency multiplier in the i.f. channel.

*Bendix Aviation Corporation. Convention date (U.S.A.) December 13th, 1945.*

627 412.—Tuning aid for f.m. receivers in which stray a.f. voltages, applied to the tuning control by the fingers, are used for amplitude modulation of the signal whereby circuit tuning is indicated by minimum response to such modulation.

*Philco Corporation. Convention date (U.S.A.) February 26th, 1946.*

627 985.—Amplitude-limiting detector circuit in which positive and negative half cycles are separately limited and combined with modulated oscillations to give a degree of asymmetry dependent on the original modulation.

*Marconi's Wireless Telegraph Co. Ltd. (assignees of M. G. Crosby). Convention date (U.S.A.) July 31st, 1945.*

### TELEVISION CIRCUITS AND APPARATUS

FOR TRANSMISSION AND RECEPTION

627 528.—Scanning-beam transmitting tube in which all but the fastest of the repelled electrons are prevented from reaching the amplifying system.

*Compagnie Pour La Fabrication des Compteurs et Materiel d'Usines A Gaz. Convention date (France) September 10th, 1946.*

627 568.—Television receiver in which the tube is in a pivoted carrier so that the viewing screen can be raised above the top of the cabinet for use.

*P.R.T. Laboratories Ltd., H. Baybut and A. J. Gale. Application date June 5th, 1946.*

627 588.—Saw-tooth generator operative only after receipt of a control pulse which is applied to a charging valve and to a valve controlling the discharging valve.

*Sadir-Carpentier. Convention date (France) May 7th, 1945.*

628 015.—Television system using time-modulated sound pulses superimposed on blanking pulses, the particular method of producing the composite signal being described.

*Standard Telephones and Cables Ltd. (assignees of N. H. Young Jr.). Convention date (U.S.A.) March 30th, 1946.*

### CONSTRUCTION OF ELECTRON-DISCHARGE DEVICES

626 289.—A cavity-resonator valve in which electrons are injected from outside the cavity wall through a grid and having a reflector electrode within an opening opposite the grid, the said electrode being insulated so that it may have a different steady potential.

*The British Thomson-Houston Company Limited, and W. J. Scott. Application date February 2nd, 1944.*

626 687.—Combined thermionic valves and waveguides for high frequencies, in which one of the waveguides is provided with two spaced shorting bridges selected in position so that the impedance of the waveguide is matched to the valve.

*The General Electric Company and E. G. James. Application date February 9th, 1944.*

627 287.—Magnetron anode structure formed by the brazed assembly of metal rings plated with a fusible metal.

*Marconi's Wireless Telegraph Co. Ltd. (assignees of L. P. Garner). Convention date (U.S.A.) May 30th, 1942.*

627 980.—Velocity-modulation valve in which the end walls of the cavities are arranged as a deformable diaphragm for tuning control.

*Compagnie Generale de Telegraphie sans Fil. Convention date (France) March 25th, 1944.*

### SUBSIDIARY APPARATUS AND MATERIALS

626 973.—A u.h.f. wavemeter embodies a resonant line tunable by a cup-shaped element adjustable around the end of an inner conductor of the line, means being included for indicating resonance.

*W. A. W. Williams and B. Solley. Application date October 5th, 1946.*

627 113.—A capacitor includes a compressed powdered ceramic dielectric whereby the capacitance can be varied by the voltage applied to the capacitor plates.

*Radio Corporation of America (assignees of H. L. Donley and C. Wentworth). Convention date (U.S.A.) April 27th, 1946.*

627 608.—Piezo-crystal circuit using several crystals of the same frequency with loss couplings between them to damp out unwanted frequencies.

*Marconi's Wireless Telegraph Company Ltd. and W. S. Mortley. Application date December 5th, 1946.*

627 870.—A band-pass filter of the resonant line type having tandem-connected parallel-disposed elements and coupling members arranged to disturb the natural coupling between said elements.

*Hazeltine Corporation (assignees of H. A. Wheeler). Convention date (U.S.A.), November 8th, 1945.*

# ABSTRACTS and REFERENCES

Compiled by the Radio Research Board and published by arrangement with the Department of Scientific and Industrial Research

*The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to it.*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## ACOUSTICS AND AUDIO FREQUENCIES

016 : 534 — 521

**References to Contemporary Papers on Acoustics.**—A. Taber Jones. (*J. acoust. Soc. Amer.*, Nov. 1949, Vol. 21, No. 6, pp. 639–646.) Continuation of 1 of January.

534.121.1 — 522

**Vibration of a Metal Plate in a Sound Field.**—T. Vogel. (*J. Phys. Radium*, July 1946, Vol. 7, No. 7, pp. 193–201.) The theory of acoustic transparency is discussed and an approximate expression is derived for the transparency of a plate for excitation frequencies considerably different from its fundamental resonance frequency. A method is described for the measurement of transparency. Results obtained for four Al plates, of thickness 1 mm and of sizes from 2 m × 1 m to 33 cm × 25 cm, confirm the theory given but are not in agreement with the theory of Davis (1933 Abstracts, p.219), especially for the smaller plates.

534.21 — 523

**General Expression for Huyghens' Principle for Attenuated Propagation of Longitudinal Waves.**—J. Brodin. (*C. R. Acad. Sci., Paris*, 14th Nov. 1949, Vol. 229, No. 20, pp. 987–991.) It is required to determine a surface distribution of sources on a closed surface S which produces the same wave outside S as a given distribution of sources inside S. If the medium is imperfectly transparent, the solution is not unique; the general solution involves a single-layer distribution and a double-layer distribution over S, one of these layers being arbitrary. Among these solutions, however, only one gives no wave inside S. See also 712 below.

534.213 — 524

**The Acoustic Characteristics of Conical Pipes.**—M. Mokhtar & G. A. Messih. (*Proc. phys. Soc.*, 1st Dec. 1949, Vol. 62, No. 360B, pp. 793–799.) Experimental investigation of slightly tapered pipes to determine end corrections and the variation of velocity and pressure along the pipe. The resonance frequencies are in fair agreement with the theoretical formula $kr = -\tan kl$ ($r$ = throat radius, $l$ = slant length and $k = \omega/c$, where $c$ is the velocity of sound). The nodes are not equidistant and pressure at the nodes is not zero. The end correction is given approximately by the empirical expression $0.6r + (0.6l - 2.100/n)\sin\theta$, where $n$ is the resonance frequency and $\theta$ the semi-apical angle of the cone.

534.231 : 534.121.1 — 525

**Pressure Distribution in the Acoustical Field Excited by a Vibrating Plate.**—J. Pachner. (*J. acoust. Soc. Amer.*, Nov. 1949, Vol. 21, No. 6, pp. 617–625.) The computed distribution is based on an exact expression for the form of the vibrations in the circular plate, and is valid at distances from the centre of the plate greater than 10–20 times its radius. A short survey of the theory of forced vibrations is included.

534.321.7 : 621.396.615.11 — 526

**Tone Source for Tuning Musical Instruments.**—E. L. Kent. (*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 164, 166.) A Hartley oscillator produces two alternative timbres for the notes A and $B_\flat$, the semitone relation being retained irrespective of the setting of a control for the frequency of A in the range 435–445 c/s.

534.321.7.08 — 527

**International Standard Musical Pitch.**—L. S. Lloyd. (*J. R. Soc. Arts*, 16th Dec. 1949, Vol. 98, No. 4810, pp. 74–85. Discussion, pp. 86–89.) A verbatim report of a paper read before the Society on 16th Nov. 1949. The history of the adoption of standard pitch is traced from the French decree of 1858 to the British Standards publication B.S.880/49 of July 1949, which defines the international standard of concert pitch to be based on a frequency of 440 c/s for the note A in the treble clef.

534.321.9 — 528

**Measurements on the Absorption of Ultrasonics in Water.**—A. van Itterbeek & P. Slootmakers. (*Physica, 's Grav.*, Oct. 1949, Vol. 15, No. 10, pp. 897–905. In English.) A balance method was used, at a frequency of 1 500 kc/s. If $\alpha$ is the absorption coefficient and $\nu$ the frequency, $\alpha/\nu^2 = 55 \times 10^{-17}$ sec$^2$/cm at 22°C, which agrees well with the value obtained for frequencies over 10 Mc/s. Absorption in a NaCl solution was also measured as a function of concentration.

534.321.9 — 529

**Scattering of an Underwater Ultrasonic Beam from Liquid Cylindrical Obstacles.**—P. Tamarkin. (*J. acoust. Soc. Amer.*, Nov. 1949, Vol. 21, No. 6, pp. 612–616.) Long summary of part of a thesis.

534.42 + 538.561 530

**Production and Observation of High Frequencies due to Combination of Several Low Frequencies.**—Dolinski. (*See* 560.)

534.612.4 531

**The Plane Wave Reciprocity Parameter and its Application to the Calibration of Electroacoustic Transducers at Close Distances.**—R. D. Simmons & R. J. Urick. (*J. acoust. Soc. Amer.*, Nov. 1949, Vol. 21, No. 6, pp. 633–635.) "The reciprocity parameter for a plane wave sound field is shown to be equal to $2A/\rho c$, where $A$ is the area of the plane piston source and $\rho c$ the characteristic acoustic resistance of the medium. It is shown experimentally that the sound field in front of a plane piston source is effectively plane over a distance approximately equal to $A/\lambda$, and that in such a region the plane wave parameter may be used to obtain a free-field calibration of the transducer by the reciprocity method."

534.78 532

**Theory of Speech Masking by Reverberation.**—R. H. Bolt & A. D. MacDonald. (*J. acoust. Soc. Amer.*, Nov. 1949, Vol. 21, No. 6, pp. 577–580.) A general statistical theory, in which speech is regarded as a series of discrete pulses distributed over a 30-db range in sound-pressure level in a given frequency band.

534.843 533

**Acoustics of Rooms.**—M. A. S. Ross. (*Nature, Lond.*, 24th Dec. 1949, Vol. 164, No. 4182, pp. 1080–1081.) Summary of a course of 6 lectures by R. H. Bolt, which included reviews of accepted theory, recent progress, and problems requiring further investigation.

534.844.4 534

**Problems of Sound in Large Halls with Strong Reverberation.**—T. S. Korn. (*HF, Brussels*, 1949, No. 4, pp. 103–108. In French, with English summary.) Reasons for reduced intelligibility in certain halls are discussed. The simplifications generally accepted for the acoustic study of halls do not apply when the reverberation time is large. Methods for reducing the reverberation effect are discussed and an 'apparent reverberation time' is introduced, in which account is taken of the duration of the exciting sound. The most effective procedure for reducing reverberation is to increase as far as possible the intensity of the direct wave relative to that of the reflected wave.

534.86 : 621.396.619.13 : 621.396.712 535

**A Demonstration of Experimental F.M. Broadcast Transmitting and Receiving Equipment.**—W. W. Honnor. (*Proc. Instn Radio Engrs, Aust.*, Feb. 1949, Vol. 10, No. 2, pp. 35–41.) 1948 Australian I.R.E. Convention paper. The equipment is described and an analysis of a listening test is given. The results indicate that f.m. broadcasting transmissions, if reproduced by high-quality receivers, will be more acceptable to the majority of listeners than the service provided by the existing a.m. m.f. system, if conventional superheterodyne receivers are used.

621.395.61.62 536

**A Low-*Q* Directional Magnetostrictive Electroacoustic Transducer.**—L. Camp & F. D. Wertz. (*J. acoust. Soc. Amer.*, Nov. 1949, Vol. 21, No. 6, p. 636.) Correction to 3022 of 1949.

621.395.61.62 537

**Theory of Passive Linear Electroacoustic Transducers with Fixed Velocity Distribution.**—L. L. Foldy. (*J. acoust. Soc. Amer.*, Nov. 1949, Vol. 21, No. 6, pp. 595–604.) In previous papers (3559 of 1945 and 1090 of 1947) a general theory of transducers was developed. The special case of a fixed velocity distribution is now analysed in simpler mathematical terms but in greater detail. The impedances, transmitting and receiving responses, directional properties, and efficiency are considered, and the 'reciprocity' and 'available acoustic power' theorems are proved. Summary noted in 939 of 1949.

621.395.61 538

**Sensitivity and Fidelity of Microphones.**—J. Henry. (*Radio franç.*, Nov. 1949, No. 11, pp. 17–20.) Two applications of the principle of reaction are first outlined and two microphone circuits then described. The first circuit comprises a microphone of any type that may act as a variable impedance element in an oscillatory circuit, a h.f. oscillator, a special type of quartz discriminator, and a frequency stabilizer. Using a capacitor microphone the l.f. output available is of the order of 1 V, compared with the normal 10–100 mV. The second arrangement is a reaction system using a carbon microphone coupled mechanically to a telephone earpiece. This can give either a considerable increase in sensitivity or in fidelity.

621.395.625 + 534.862.3/.4 539

**Some Factors governing the Choice of a High-Quality Recording and Reproducing System.**—G. F. Dutton. (*Proc. Instn Radio Engrs, Aust.*, Oct. 1949, Vol. 10, No. 10, pp. 269–275.) A general discussion of the advantages and limitations of film recording, the Philips-Miller system (mechanically engraved variable-area film), disk systems and magnetic-tape systems.

621.395.625.2 540

**Modern Practice in Disc Recording.**—R. V. Southey. (*Proc. Instn Radio Engrs, Aust.*, Nov. 1948, Vol. 9, No. 11, pp. 16–20.) A general survey, with special reference to needle-tip and groove dimensions, cross-over frequency and frequency correction. Recommended tip and groove dimensions for standardization are noted.

621.395.625.2 : 534.86 541

**Some Problems of Disc Recording for Broadcasting Purposes.**—F. O. Viol. (*Proc. Instn Radio Engrs, Aust.*, Feb. 1949, Vol. 10, No. 2, pp. 42–47.) The advantages of disk recording are outlined and the best compromise for groove spacing and velocity and direction of cut for 16-in. disks is discussed. A recommended recording characteristic for general use is given. Recording heads and pickups suitable for cellulose-nitrate disks are also considered.

621.396.619.23 : 621.396.615.11 542

**Warbler for Beat-Frequency Oscillator.**—Flanagan. (*See* 590.)

621.396.645 543

**Description and Analysis of a New 50-Watt Amplifier Circuit.**—McIntosh & Gow. (*See* 592.)

621.396.645.37 544

**The Cathode Follower as Audio Power Amplifier.**—Sterling. (*See* 599.)

681.84 545

**Designing a Portable Sound Projector.**—G. A. del Valle. (*Elect. Mfg, N.Y.*, Aug. 1949, Vol. 44, No. 2, pp. 74–79, 188.) The R.C.A.400 Junior model, for use with 16-mm sound film, weighs only 46 lb with its loudspeaker. The sound amplifier has four stages, with inverse feedback, and gives up to 10 W sound output with $<2\%$ distortion, while frequency response varies by less than 5 db over the range 50–5 000 c/s. An intensity of 150 lumens is obtained on the screen, using a standard 1 000-W lamp with forced-air cooling. The apparatus gives a class-room performance comparable with professional cinema standards and it does not require a skilled operator.

681.85 546

**Some Aspects of Phonograph Pickup Design.**—W. R. Nicholas. (*Proc. Instn Radio Engrs, Aust.*, March 1949, Vol. 10, No. 3, pp. 63–73. Correction, Vol. 10, No. 4, p. 113.) A brief review is given of the properties of lateral disk records. The electrical and mechanical requirements to be met by a pickup are stated and a general treatment of the design of a typical magnetic pickup is given. The forces likely to be encountered at the stylus point are discussed and calculations are made for a particular type of magnetic pickup. Improved types are briefly described and their principal characteristics noted.

## AERIALS AND TRANSMISSION LINES

621.315.212 : [621.395 + 621.397 547

**Progress in Coaxial Telephone and Television Systems.**—Abraham. (*See* 739.)

621.392.22 + 621.392.26† 548

**A General Method of Solution for Non-Uniform Cylindrically Symmetrical Wave Fields.**—H. H. Meinke. (*Z. angew. Phys.*, Oct. 1949, Vol. 1, No. 11, pp. 509–516.) Continuation of 1645 of 1948. The frequency dependence of the permeability $\mu$ of the dielectric in the transformed system is determined by the Fourier coefficients in the expression for $\mu$. To a first approximation the permeability increases as the square of the frequency. The wave field of a cylindrically symmetrical non-uniformity can be determined as soon as the e.s. field distribution for the transformed system and its equipotential lines are known.

621.392.22 : 621.3.09 549

**Contribution to the Study of Propagation on a Heterogeneous Line.**—F. H. Raymond. (*J. Phys. Radium*, June 1946, Vol. 7, No. 6, pp. 171–177.) Methods of solving the equations of propagation, due to Parodi (2149 of 1944) and to Raymond (2472 of 1946), are discussed. The coefficients of reflection and transmission are defined for the general case, and a solution is obtained by means of series. Steady-state conditions are considered; Laplace transformations can be used to apply the results obtained to transient conditions. The method of treatment is applicable to coaxial cables with irregularities which may give rise to echo signals, and to propagation in heterogeneous media.

621.392.26† 550

**The Field in Non-Uniform Rectangular Waveguides with $H_{10}$-Wave Excitation.**—R. Piloty, Jr. (*Z. angew. Phys.*, Oct. 1949, Vol. 1, No. 11, pp. 490–502.) Continuation of 31 of January. An approximation method of solution of the general equation is developed. See also 548 above.

621.392.26† 551

**On the Input Impedance of a Waveguide beyond its Cut-Off Frequency.**—A. Briot. (*C. R. Acad. Sci., Paris*, 21st Nov. 1949, Vol. 229, No. 21, pp. 1066–1068.) Bromwich's differential equation for the e.m. field can be simplified in the case of a piston attenuator, the excitation being longitudinal and periodic. Formulae are obtained for the impedance for both electric and magnetic modes. The input impedance of a piston attenuator is imaginary, whatever the mode considered. It is capacitive for electric modes and inductive for magnetic modes. Since the characteristic impedance of a waveguide is a real quantity, a dissipative element should be placed in front of the attenuator to absorb reflected energy.

621.392.26† : 621.392.43 : 621.396.615.141.2 552

**Determination of the Characteristics of Matching Circuits for a Magnetron Modulator.**—J. Ortusi & P. Fechner. (*Ann. Radioélect.*, Oct. 1949, Vol. 4, No. 18, pp. 295–314.) The characteristic impedances of waveguides are calculated for different types of wave, and the conditions for suppression of the stationary waves caused by the insertion of an active obstacle in the waveguide are considered in detail. A method is described for measurement of the reflection, transmission and loss characteristics of an active obstacle in a waveguide and, for the case of a magnetron modulator, the conditions are determined which are necessary for the greatest possible variation of the impedance thrown back by the magnetron into the waveguide. See also 804 of 1949 (Gutton & Ortusi) and 2140 of 1949.

621.396.67 553

**Fundamental Limitations of Small Antennas.**—H. A. Wheeler. (*Proc. Instn Radio Engrs, Aust.*, Feb. 1949, Vol. 10, No. 2, pp. 47–52.) Reprint. See 1261 of 1948.

621.396.67 554

**Radiation Patterns and Gain of a Four-Antenna Array located at the Corners of a Square around a Central Parasitic Antenna.**—G. Boudouris. (*J. Brit. Instn Radio Engrs*, Dec. 1949, Vol. 9, No. 12, pp. 427–439.) Analysis assuming all are $\lambda/4$ aerials grounded to a perfectly conducting plane. The impedance at the base of each aerial is calculated as a function of the length of the diagonal, and the horizontal and vertical radiation patterns are determined. These may have directional characteristics. By comparing the field produced by the array to the field produced by a single aerial for the same power input, the average increase in field is found as a function of the diagonal spacing. The maximum average increase is 20%, and the maximum peak increase in certain directions 60%. The effect of limited ground conductivity and the use of radiators shorter than $\lambda/4$ are discussed.

621.396.67 555

**Helical Beam Antenna Design Techniques.**—J. D. Kraus. (*Communications*, Sept. 1949, Vol. 29, No. 9, pp. 6–9, 35.) Approximate formulae are derived for beam width, power gain, axial ratio and terminal resistance of a helix radiating in the axial mode with angle of pitch between 12° and 15°. The smaller angle gives a sharper main lobe for given axial length, while the larger angle gives a slightly smaller resistance variation throughout the frequency band.

621.396.67 : 621.392.43 556

**Impedance Transformation in Folded Dipoles.**—R. Guertler. (*Proc. Instn Radio Engrs, Aust.*, April 1946, Vol. 10, No. 4, pp. 95–100.) The impedance of a folded dipole, relative to that of a simple dipole, may be adjusted by the use of conductors of different diameters for the separate elements. Increased impedance ratios can be obtained by the use of additional elements. Approximate formulae are developed for calculating the impedance ratio when the current ratio is known. Measurements confirm the practical applicability of the formulae.

621.396.67 : 621.397.62 557

**Quadrature Phased TV Receiving Antenna.**—(*Tele-Tech*, Nov. 1949, Vol. 8, No. 11, pp. 36–37.) A capacitive phase shifter, with combined resistive and inductive detuning compensation, is used with two crossed horizontal dipoles to provide variable directivity without rotation of the aerial system. Orientation of one dipole can be used to minimize ghost effects. An outdoor and a built-in system are described. The whole U.S. television band is covered without switching.

621.396.671 558

**The Predetermination of Antenna Characteristics by Means of Models.**—R. D. Boadle. (*Proc. Instn Radio*

*Engrs, Aust.*, June 1949, Vol. 10, No. 6, pp. 155–159.) Paper presented at the Australian I.R.E. convention, Sydney, 1948. Description of equipment and technique, with results for a high-gain vertically polarized v.h.f. aerial.

559

621.396.67 : 621.396.9

**Radar Scanners and Radomes.** [Book Review]—Cady, Karelitz & Turner. (*See* 651.)

## CIRCUITS AND CIRCUIT ELEMENTS

560

534.42 : 537.591

**Production and Observation of High Frequencies due to Combination of Several Low Frequencies.**—S. Dolinski. (*C. R. Acad. Sci., Paris*, 24th Oct. 1949, Vol. 229, No. 17, pp. 812–814.) Suitable methods are described for audio and radio frequencies.

561

621.3.015.33 : 621.392

**Response of Circuits to Steady-State Pulses.**—D. L. Waidelich. (*Proc. Inst. Radio Engrs*, W. & E., Dec. 1949, Vol. 37, No. 12, pp. 1396–1401.) "A method of calculating the steady-state response of circuits to repeated pulses is given using the method of the steady-state operational calculus. A short table of transforms which have been found useful in these calculations is also presented. The response of several basic circuits to these pulses is obtained and shown as calculated curves, and the calculated curves are then compared with curves obtained experimentally. These curves have been found to be very useful in adjusting circuits to be used with pulses. Several other possible applications are discussed."

562

621.314.2 : 621.396.611.33/.34

**A Design for Double-Tuned Transformers.**—J. B. Rudd. (*Proc. Instn Radio Engrs, Aust.*, Jan. 1949, Vol. 10, No. 1, pp. 3–9.) 1948 Australian I.R.E. Convention paper. See 3063 of 1949.

563

621.314.3*

**An Experimental Study of the Magnetic Amplifier and the Effects of Supply Frequency on Performance.**—E. H. Frost-Smith. (*J. Brit. Instn Radio Engrs*, Dec. 1949, Vol. 9, No. 12, pp. 440–443.) Discussion on 44 of January.

564

621.314.3*

**On the Dynamics of the Self-Excited Series Transductor.**—M. Delattre & K. Kühnert. (*C. R. Acad. Sci., Paris*, 24th Oct. 1949, Vol. 229, No. 17, pp. 819–821.) Discussion for the period of transition from one steady state to a different one when the control current is changed. The response time depends on the absolute value of the input signal and on various other factors; it approaches zero under ideal conditions.

565

621.314.3*

**Self-Saturation in Magnetic Amplifiers.**—W. J. Dornhoefer. (*Elect. Engng, N.Y.*, Nov. 1949, Vol. 68, No. 11, p. 988.) Summary of A.I.E.E. paper. A high voltage can be induced in the control winding if one a.c. coil becomes inoperative and the effective inductance and time delay of the control circuit are substantial, but these factors can be reduced either by external feedback or by self-saturation. The latter method makes use of a rectifier valve in series with each a.c. (anode) winding; its advantages are discussed. It has been applied for both a.c. and d.c. output and for single and polyphase supplies. Rectifiers with high backward/forward resistance ratio are required.

566

621.317.772

**Diode Phase-Discriminators.**—R. H. Dishington. (*Proc. Inst. Radio Engrs*, W. & E., Dec. 1949, Vol. 37, No. 12, pp. 1401–1404.) "Two sinusoidal phase-discriminators are analyzed and it is found that universal curves of their general phase characteristics can be plotted as a function of two parameters. From these curves it is concluded that the resistances in series with the tubes and also the tube resistances themselves are the most important factors in determining optimum performance."

567

621.318.4 : 621.397.62

**Coils for Television Receivers.**—F. Juster. (*Télévis. franç.*, Nov. 1949, No. 53, pp. 15–18, 24.) Graphical methods for designing h.f., i.f. and video-frequency coils with maximum *Q*.

568

621.318.42

**On the Calculation of Effective Core Permeability of Premagnetized Choke Coils.**—J. Kammerloher. (*Funk u. Ton*, 1949, Vol. 3, Nos. 9/10, pp. 491–496.) Continuation of 29 of 1949. Values obtained when differential permeability is replaced by reversible permeability are in good agreement with those of Feldtkeller.

569

621.318.7

**The Theory of Electric Filters and the Polynomials of Tchebycheff.**—A. Colombani. (*J. Phys. Radium*, Aug. 1946, Vol. 7, No. 8, pp. 231–243.) Full paper for abstract of shorter version see 652 of 1947.

570

621.392

**Quarter Wave Networks.**—E. Green. (*Marconi Rev.*, Oct./Dec. 1949, Vol. 12, No. 95, pp. 157–171.) The properties of a normal $\lambda/4$ line are derived by means of vector diagrams and the following basic equations used to express any form of $\lambda/4$ or inverting network:

$$I_1 = jV_2/R_0, \quad V_1 = jR_0I_2.$$

$I_1$, $V_1$ and $I_2$, $V_2$ are respectively the input and output current and voltage and $R_0$ is the surge impedance of the $\lambda/4$ line. It is proved that certain types of lumped network are identical as regards external relations. The component values and characteristic impedances of various balanced and unbalanced networks are tabulated. An appendix deals with the derivation of different forms of a bridge network devised by A. T. Starr for operation between balanced and unbalanced lines. This derivation is based on that given by Starr in an unpublished memorandum.

571

621.392.43 : 621.396.67

**Design Procedures for Pi-Network Antenna Couplers.**—L. Storch. (*Proc. Inst. Radio Engrs*, W. & E., Dec. 1949, Vol. 37, No. 12, pp. 1427–1432.) "The design of reactive pi-networks for transforming a wide range of complex load impedance into a fixed resistance shunted by a tuned circuit is subjected to a thorough investigation. A very significant result is the complete analogy which is established between the analysis and design of the pi-network and the equivalent manipulation of a group of simple geometrical figures."

572

621.392.5

**Realization of Linear Quadripoles with Prescribed Frequency Dependence, taking account of Equal Coil and Capacitor Losses.**—Nai-Ta Ming. (*Arch. Elektrotech.*, 1949, Vol. 39, No. 8, pp. 496–507.) See also 3367 of 1949.

573

621.392.5

**Resistance Quadripoles as Attenuator Elements.**—W. Taeger. (*Funk u. Ton*, 1949, Vol. 3, Nos. 9/10, pp. 475–487.) Calculations of various impedance ratios in T, L, Π and bridged-T networks. Attenuation in nepers is tabulated and plotted for different values of R/Z.

574

621.392.5

**On the Synthesis of the Most General Passive Quadripoles.**—V. Belevitch. R. Leroy. (*Câbles & Transmission, Paris*, Oct. 1949, Vol. 3, No. 4, pp. 340–341.) Comment on 1886 of 1949 and author's reply.

621.392.52 575

**General Forms of Ladder-Filter Half-Sections Classed according to the Value of the Image-Impedance Transfer Index.**—J. E. Colin. (*Câbles & Transmission, Paris*, Oct. 1949, Vol. 3, No. 4, pp. 281–293.) Continuation of 2748 of 1949. Different simple and compound types of ladder-filter network with one or two cut-off frequencies are reduced to eight basic forms, whose principal characteristics are discussed. Formulae for impedance ratio, component values and image-impedance are tabulated, and three general laws of form are enunciated.

621.392.52 576

**Transference Nomographs for Low-Pass Iterative Filters.**—E. W. Tschudi. (*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 112, 114.) For determining time constant and phase lag, given input frequency, attenuation and number of stages up to 6.

621.392.52 577

**Using the Reactance Chart for Filter Design Problems.**—H. B. Davis. (*Audio Engng, N.Y.*, Dec. 1949, Vol. 33, No. 12, pp. 12–13, 40.) Methods are outlined for determining the values of components of various filter and equalizer circuits.

621.392.52 578

**Filters: Part 1 — The Importance of $Z_0$, the Characteristic Impedance.**—"Cathode Ray". (*Wireless World*, Jan. 1950, Vol. 56, No. 1, pp. 25–29.)

621.392.52 : 621.395.44 : 621.316.1 579

**Wave Filters protect Carrier Signals from Shunting Effect of Capacitors.**—W. A. Ringger, Jr. (*Elect. World, N.Y.*, 19th Nov. 1949, Vol. 132, No. 21, pp. 106–108.) High-voltage wave filters using various parallel and series resonant-circuit combinations preserve carrier voltages and suppress spurious voltages associated with carrier operation on distribution systems.

621.395.667 580

**Equalizer Charts.**—(*Bell Syst. tech. Publ. Monogr.* B-1643, 15 pp.) Loss and phase of series, shunt and bridged-T equalizers are given in decibels and degrees.

621.396.611 581

**Contribution to the Theory of the Mathematical Treatment of Nonlinear Phenomena.**—H. Rosenhamer. (*Bull. schweiz. elektrotech. Ver.*, 8th Jan. 1949, Vol. 40, No. 1, pp. 5–18. In German.) A method is given for the solution of the general equation of order $n$ applicable to nonlinear electrical phenomena. The equation is reduced to the sum of products of linear differential quotients with functional coefficients, which are then replaced successively by suitably chosen constants and the integration carried out by normal methods. The solution thus obtained only satisfies the original equation approximately, but can be used to determine fresh values of the functional coefficients giving a better solution. The method is applied to the study of an LCR circuit to which a sinusoidal voltage is applied.

621.396.611.1 582

**Some Theoretical Considerations and Experiments on Oscillation Phenomena in Circuits with Nonlinear Elements.**—G. J. Elias & S. Duinker. (*Tijdschr. ned. Radiogenoot.* Nov. 1949, Vol. 14, No. 6, pp. 163–191. In Dutch, with English summary.) The solutions of the equation

$$\ddot{y} + y + \nu y^3 = \alpha \cos(px + \phi)$$

are discussed. They include periodic solutions giving a fundamental frequency which is a submultiple of the generator frequency.

For free vibrations, the period can be expressed as an elliptic integral, for forced vibrations, an extension of this expression is given and the stability of the solution is discussed. If the e.m.f. is varied continuously, several discontinuities must occur if a periodic forced vibration, of the fundamental or a subharmonic frequency, is to be sustained. The effect of adding any number of odd powers of $y$ to the left hand side of the differential equation is also considered; the expression for the period then becomes a hyperelliptic function. Experiments which confirm the theory are described; oscillograms of subharmonic vibrations are shown.

621.396.611.1 583

**Resonant Circuits with Time-Varying Parameters.**—R. H. Kingston. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1478–1481.) An approximate solution of the differential equation, with an error criterion giving the limits of accuracy of the solution.

621.396.611.4 584

**Analogue Studies of Losses in Reflex Oscillator Cavities.**—F. W. Schott & K. R. Spangenberg. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1409–1418.) "An analysis is made which shows the method of applying the network analogue to investigation of the effect of dielectric and wall losses on cavity-resonator behavior.

"The $Q$ and shunt resistance of re-entrant cavities operating in the first- and second-order $TM_0$ type modes are investigated. The condition for a zero of shunt resistance is determined. Experimental results are discussed."

621.396.615 585

**The Breadth of the Pulling-In Range of a Self-Excited Valve Generator controlled by an Integral Multiple of its Natural Frequency.**—H. Kanberg. (*Funk u. Ton*, 1949, Vol. 3, Nos. 9/10, pp. 495–505.) Analysis showing how the frequency range may be determined from the Lissajous figure displayed on a c.r.o.

621.396.615 586

**The Transitron Effect and its Applications.**—J. Moline. (*Radio franç.*, Nov. 1949, No. 11, pp. 21–24.) Distinction is made between the dynatron effect in a tetrode and the transitron effect in a pentode, and a simple explanation of transitron operation is given. The characteristics of some typical valves operating as transitrons are noted, and applications of the transitron effect to the maintenance of sinusoidal or of relaxation oscillations are considered.

621.396.615 : 621.396.611.32 587

**Phase-Shift Oscillators with Very Tight Coupling.**—M. Soldi. (*Radio tech. Dig., Edn franç.*, Dec. 1949, Vol. 3, No. 6, pp. 353–363.) French version of 2757 of 1949, with the addition of a comprehensive bibliography.

621.396.615.17/.18 588

**High-Ratio Multivibrator Frequency Divider.**—M. Silver & A. Shadowitz. (*Proc. Instn Radio Engrs, Aust.*, Sept. 1949, Vol. 10, No. 9, pp. 256–258.) Reprint. See 52 of 1949.

621.396.615.17 : 621.317.755 589

**A Slow-Sweep Time-Base.**—V. H. Attree. (*J. sci. Instrum.*, Aug. 1949, Vol. 26, No. 8, pp. 257–262.) The requirements of timebases suitable for sweep durations of the order 100 sec–10 ms are discussed with special reference to the use of triggered multiple sweeps. Flyback time can be made as little as one thousandth of the scan time. The circuit described is direct-coupled throughout, and linear traces of identical velocity and duration are generated irrespective of the nature of the triggering wave. The timebase is unaffected by any trigger pulses which may occur during sweeping.

621.396.619.23 : 621.396.615.11 590

**Warbler for Beat-Frequency Oscillator.**—J. L. Flanagan. (*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 93–95.) A new reactance valve circuit giving f.m. of about ± 10% without accompanying a.m., for use with a standard beat-frequency audio oscillator. Instead of using a capacitor to couple anode and grid, the grid is earthed and the capacitor is connected between anode and cathode, a degenerative resistance being introduced between cathode and earth. The theoretical expression for the input impedance shows that its resistive component is independent of $g_m$.

621.396.645 591

**A Contact-Modulated Amplifier and Some of Its Laboratory Uses.**—J. F. Lash. (*Science*, 7th Oct. 1949, Vol. 110, No. 2858, pp. 374–375.) A d.c. signal, converted to 80-c/s a.c. by a mechanical interrupter, is stepped up in voltage by a transformer followed by several valve-amplifier stages and finally rectified by a second interrupter synchronized with the first. This system enables d.c. signals of a few hundredths of a microvolt to be measured.

621.396.645 592

**Description and Analysis of a New 50-Watt Amplifier Circuit.**—F. H. McIntosh & G. J. Gow. (*Audio Engng, N.Y.*, Dec. 1949, Vol. 33, No. 12, pp. 9–11 . . 40.) A circuit is described which eliminates the distortion present in conventional push-pull amplifiers operating between class A and class B, due to leakage reactance between the primary windings. A high ratio (>200 000 to 1) of primary inductance to leakage reactance is obtained by winding the two primaries together in a bifilar manner. Circuit details and performance characteristics are given. A 50-W output is obtained at frequencies from 20 c/s to 20 kc/s with less than 1% harmonic or intermodulation distortion.

621.396.645 593

**Grounded-Grid Power Amplifiers.**—P. A. T. Bevan. (*Radio tech. Dig., Edn franç.*, Dec. 1949, Vol. 3, No. 6, pp. 323–339.) French version of 2763 of 1949, with the addition of a comprehensive bibliography.

621.396.645 594

**Wide-Band Chain Amplifier.**—F. Kennedy & G. Rudenberg. (*Elect. Mfg, N.Y.*, Nov. 1949, Vol. 44, No. 5, pp. 56–59 . . 168.) The amplifier operates on the travelling-wave principle; a series of valves have their grids connected to one delay line and their anodes to another. The forward waves produced in the lines augment each other and the backward waves are absorbed by the anode resistors. Six valves are used in an amplifier with a constant gain of 20 db over the range 200 kc/s–200 Mc/s. Undistorted output is obtained up to 4 V and the s.w.r. is < 1.5 db. Design and construction details are given; the importance of careful layout to avoid stray capacitance and inductance is stressed.

621.396.645 : 621.385.029.63/.64 595

**200-Mc/s Traveling-Wave Chain Amplifier.**—H. G. Rudenberg & F. Kennedy. (*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 106–109.) A brief descriptive account of a wide-band amplifier consisting of several valves whose grids are fed from taps on a terminated artificial line consisting of a series of inductors connected between the grid shunt capacitors. The output is taken from the end of a similar line joining the anodes. Particular examples and some refinements are discussed. The optimum gain per chain is 2.718. Any further gain should be obtained by connecting such chains in series. Such a design is most economical in valves.

621.396.645.029.42/.52 : 621.396.822 596

**Selective Amplification at Low Frequencies.**—L. de Queiroz Orsini. (*Onde élect.*, Nov. & Dec. 1949, Vol. 29, Nos. 272 & 273, pp. 408–413 & 449–456.) Discussion of voltage amplifiers with narrow pass-band, for frequencies between 2 c/s and 50 kc/s. Feedback circuits are described for frequencies near 500 c/s, and parallel-T circuits for very low frequencies. Flicker effect and other valve and circuit noise, and methods of reducing it, are also considered.

621.396.645.211 597

**Theory of Resistance-Amplifier Stage.**—W. Druey. (*Bull. schweiz. elektrotech. Ver.*, 22nd Jan. 1949, Vol. 40, No. 2, pp. 49–51. In German.) A derivation of the gain formula from a consideration of equivalent circuits.

621.396.645.35 598

**Stabilizing a Wide-Band D.C. Amplifier.**—A. J. Williams, Jr, W. G. Amey & W. McAdam. (*Elect. Engng, N.Y.*, Nov. 1949, Vol. 68, No. 11, p. 934.) Summary of A.I.E.E. paper. The problem of zero stabilization in d.c. amplifiers with negative feedback is discussed. A comparison of input with output divided by gain can be used to indicate zero disturbance, whether a signal is present or not. The comparison voltage is applied to a correcting circuit using a contact-modulated d.c. amplifier and integrator. Zero correction to within a fraction of a microvolt is thus obtained in less than 0.005 sec.

621.396.645.37 599

**The Cathode Follower as Audio Power Amplifier.**—H. T. Sterling. (*Audio Engng, N.Y.*, Dec. 1949, Vol. 33, No. 12, pp. 14–15 . . 29.) As an output stage, the cathode follower is a feedback amplifier with special characteristics which can be provided equally well by conventional amplifier circuits.

621.396.662.029.64 : 621.392.26? 600

**Microwave Attenuators.**—R. Malvano. (*Elettronica & Televisione, Turin*, Sept. 1949, Vol. 4, No. 6, pp. 221–226; French version in *Radio tech. Dig., Edn franç.*, Dec. 1949, Vol. 3, No. 6, pp. 369–374.) A general discussion with particular reference to waveguide attenuators. A description of an experimental attenuator made at the Istituto Elettrotecnico Nazionale Galileo Ferraris, Turin, is included.

## GENERAL PHYSICS

535.3 601

**On the Attenuation of Plane Waves by Obstacles of Arbitrary Size and Form.**—H. C. van de Hulst. (*Physica, 's Grav.*, Sept. 1949, Vol. 15, Nos. 8/9, pp. 740–746. In English.) The extinction cross-section of any obstacle is $-2\lambda$ times the imaginary part of the amplitude function for forward scattered light. This relation holds for any size, shape or composition of the obstacle, and may be applied to light, sound and electron scattering. It may be generalized to include polarization effects. Examples are discussed.

537.311.62 : 621.318.4 602

**Study of the H.F. Resistance of a Stranded-Wire Coil.**—A. Colombani. (*J. Phys. Radium*, Oct. 1949, Vol. 10, No. 10, pp. 285–294.) Starting from Maxwell's equations and assuming the radius of the strand to be less than the depth of penetration of the current, a simple formula is derived for h.f. resistance. The formula is in agreement with the calculations of Sommerfeld and with measured values. Its application to different problems is illustrated and shows that there is an optimum resistance and form of coil in each case. For frequencies above 3 Mc/s, solid wire is preferable to stranded wire, for which the losses in the enamel insulation of the strands become important.

537.533 : 538.569 : 621.396.822 **603**

**Fluctuation Phenomena arising in the Quantum Interaction of Electrons with High-Frequency Fields.**—D. K. C. MacDonald & R. Kompfner. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1424–1426.) Quantum-mechanical analysis of the interaction of an electron beam with an oscillating resonant cavity is applied to determine the fluctuations in energy flow in the beam. Comparison of the expressions for the classical and quantal cases indicates that, under extreme *limiting* conditions of operation, a difference might just be perceptible.

537.581 : 621.385.032.213 **604**

**Thermionic Electron Emission from Carbon.**—H. F. Ivey. (*Phys. Rev.*, 15th Aug. 1949, Vol. 76, No. 4, p. 567.) Values of the electronic work function $\phi$, the Richardson constant A, and the emission current density are given for a carbon filament after heating to 2 500°K. $\phi = 4.60$ eV and $A = 46$ A/cm² degree². The values of current density at 1 500°K and 2 000°K are 0.039 $\mu$A/cm² and 0.50 mA/cm² respectively, which are lower than the values found by other investigators.

538.3 **605**

**The Application of Maxwell's Second Law.**—H. Kafka. (*Elektrotechnik, Berlin*, Nov. 1949, Vol. 3, No. 11, pp. 353–358.) A distinction between the mathematical expression and the physical interpretation of the laws of induction.

538.51 **606**

**Electromagnetic Induction in a Rotating Sphere.**—E. C. Bullard. (*Proc. roy. Soc. A*, 7th Dec. 1949, Vol. 199, No. 1059, pp. 413–443.) An account is given of the induction of electric currents in a rotating, conducting sphere surrounded by and in contact with a concentric, stationary, conducting shell of any thickness. The currents and associated mechanical couples induced by various constant fields are discussed, and the freely decaying modes are studied. The modes involving toroidal magnetic fields are of most interest and may be important in terrestrial magnetism. Induction in an oscillating sphere is briefly discussed.

538.56 : 537.71 **607**

**On the Resistance of Electromagnetic Waves.**—C. Budeanu. (*Rev. gén. Elect.*, Nov. 1949, Vol. 58, No. 11, pp. 481–484.) This resistance could be given different values according as the classical or the rationalized system of units is used. The precautions necessary in expressing this quantity and the value it should be given are considered.

538.566 : [537.562 + 537.525.92 **608**

**Wave Amplification by Interaction with a Stream of Electrons.**—J. A. Roberts. (*Phys. Rev.*, 1st Aug. 1949, Vol. 76, No. 3, pp. 340–344.) Discussion of the propagation of plane e.m. waves in a uniform ionized medium in which the electrons have a mean drift velocity. The treatment is based on Bailey's general theory (2785 and 3406 of 1949). For frequencies below a critical frequency of the same order as the electron plasma frequency, one of the eight waves possible in general grows in amplitude as it progresses through the medium. The relation of this result to theories of solar noise and of the travelling-wave amplifier is discussed.

538.569.4 **609**

**Absorption of Ultra High-Frequency Radio Waves in Organic Liquids.**—S. C. Sirkar & S. N. Sen. (*Nature, Lond.*, 17th Dec. 1949, Vol. 164, No. 4181, pp. 1048–1049.) Acetone shows an absorption peak at about 400 Mc/s and glycerine at about 500 Mc/s for a temperature of 31°C. Absorption curves for these and other liquids are given and discussed.

## GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

016 : 550.3 **610**

**List of Recent Publications.**—H. D. Harradon. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 308–314.) Subjects covered include terrestrial magnetism and electricity, cosmic rays, upper-air research, and the earth's crust and interior.

061.3 : 551.510.535 : 538.566 **611**

**The Conference on Ionospheric Research.**—(*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 281–294.) Summaries of symposia and abstracts of papers presented at the conference held at Pennsylvania State College under joint sponsorship of that College and of the Geophysical Research Directorate, U.S.A.F., 27th–29th June 1949. Selected papers are noted below, in this section or in the 'Propagation of Waves' section.

523.32 : 621.396.822 **612**

**Microwave Thermal Radiation from the Moon.**—J. H. Piddington & H. C. Minnett. (*Aust. J. sci. Res., Ser. A*, March 1949, Vol. 2, No. 1, pp. 63–77.) Measurements have been made in a 15-Mc/s band centred at 24 kMc/s. The average temperature over the lunar disk varies sinusoidally between 198.7°K and 279.3°K, with a phase lag behind the lunar phase angle of about 45°. The discrepancy with previous measurements at long infra-red wavelengths is explained in terms of radiation from subsurface layers which are partially transparent to 24-kMc/s waves. Results are consistent with the existence of a thin layer of dust covering a solid lunar surface. The temperature of the disk of the new moon is estimated at 156°K, and that of the deep interior at 241°K.

523.53 : 621.396.9 **613**

**A Study of Radio Reflections from Meteor Trails in Research on the Upper Atmosphere.**—A. G. McNish. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, p. 285.) Summary of Pennsylvania State College Conference paper. Discussion of radio research on meteors in progress at C.R.P.L., with special reference to the investigation of ionic processes in the upper atmosphere. Experimental equipment consists of two 10-kW pulse transmitters operating continuously on 27.2 and 40.4 Mc/s. Signals reflected from the ionized trail left by the meteor are recorded on pen-and-ink tape and on photographic film. Results already obtained indicate clearly that meteors are not responsible for most of the $F_2$ reflections ordinarily obtained by vertical-incidence recorders. Two possible mechanisms of reflection are considered and an experimental method for discriminating between them is suggested.

523.72 + 523.854] : 621.396.822 **614**

**Solar and Galactic R.F. Radiation.**—M. Laffineur. (*Onde élect.*, Nov. 1949, Vol. 29, No. 272, pp. 402–407.) A general survey of recent work; results obtained by various authors are compared graphically. A solar disturbance observed at Meudon on 26th March 1949 is discussed. Other such disturbances were noted in 608 and 3129 of 1949. See also 2777 of 1948 (Lehmann).

523.72 : 621.396.11 **615**

**Observations of the Propagation of Broadcasting Waves, $\lambda \cong 1250$ m, during the Occurrence of Mögel-Dellinger Effects.**—Lauter. (*See* 716.)

523.72 : 621.396.822 **616**

**Solar Radiation at 1 200 Mc/s, 600 Mc/s and 200 Mc/s.**—F. J. Lehany & D. E. Yabsley. (*Aust. J. sci. Res., Ser. A*, March 1949, Vol. 2, No. 1, pp. 48–62.) Daily observations made between 18th Aug. and 30th Nov. 1947 are described. The characteristics of the radiation at 200 Mc/s were in general agreement with other observations. At 600 and 1 200 Mc/s the received intensity

showed long-term variations over a range of about 2 : 1. " The radiation received when the sun was almost free of sunspots corresponded to an effective black-body temperature of 0.5 million °K at 600 Mc/s and 0.2 million °K at 1 200 Mc/s. As sunspots appeared, the temperature rose and showed marked correlation with sunspot area. It is considered that radiation at these frequencies is entirely thermal in origin and that the long-period variations are at least partly due to the influence of the magnetic field of sunspots on the mechanism of thermal emission from a magneto-ionic medium. On a few occasions, isolated disturbances were observed on 600 Mc/s and 1 200 Mc/s, some of which were associated with chromospheric flares and radio fade-outs. The difficulties arising in the calibration of the apparatus and the steps taken to overcome them are discussed in detail."

523.746 : 523.752 617

**Hall Currents and the Ejection of Prominences by Sunspots.**—R. G. Giovanelli. (*Aust. J. sci. Res., Ser. A*, March 1949, Vol. 2, No. 1, pp. 39–47.) A theoretical explanation of the occurrence of eruptive prominences in the neighbourhood of a chromospheric flare. Localized regions of large space charge occur because of the relatively low conductivity *perpendicular* to lines of magnetic force. Large electric fields are thus set up normal to the magnetic field, and both positive and negative charges drift in the same sense in a direction at right angles to the crossed fields, giving rise to a general movement of the gas. The drift velocities appear to be comparable with those found in prominences. See also 78 and 376 of 1949 and 357 of February.

523.746"1949.04/.06" 618

**Provisional Sunspot-Numbers for April to June 1949.**—M. Waldmeier. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, p. 300.)

523.75 : 551.510.535 619

**Ionospheric Effects of Solar Flares, 1948.**—O. E. H. Rydbeck & D. Stranz. (*Chalmers tekn. Högsk. Handl.*, 1949, No. 83, 16 pp. In English.) Regular recordings of such effects with different kinds of apparatus were started at the Chalmers University Geophysical Observatory early in 1948. The results of the first half-year are discussed in this preliminary report. The statistical distribution across the solar disk of radiation sources giving rise to a greater or less degree of fading of radio signals is shown. The magnitude and probability of the absorption of the ultraviolet fade-out radiation in the solar corpuscular beam is discussed.

550.38"1949.01/.03" 620

**International Data on Magnetic Disturbances, First Quarter, 1949.**—(*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 295–299.) Continuation of 3418 of 1949 in revised form, with explanation of new symbols and abbreviations. The original *K*-index is replaced by a planetary 3-hour range index $K_p$. Additional preliminary data on sudden commencements and solar flare effects (crochets) are included.

550.38"1949.04/.06" 621

**Cheltenham** [Maryland] **Three-Hour-Range Indices *K* for April to June, 1949.**—P. G. Ledig. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, p. 300.)

550.384 622

**The *K*-Index of Geomagnetic Activity at Eskdalemuir, 1940–47.**—J. Crichton. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 275–276.) Discussion of the relative importance of horizontal intensity H, declination D, and vertical intensity V as the determining factor in the assignment of *K*-indices, and of the diurnal and seasonal incidence, and recurrence tendencies, of occasions when D or V was the determining factor.

550.385"1949.01/.06" 623

**Principal Magnetic Storms** [Jan.–June 1949].—(*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 301–302.)

551.510.5 : 551.524.7 624

**Tropopause Fluctuations due to High-Reaching Influxes of Cold Air.**—W. Hesse. (*Z. Met.*, May/June 1949, Vol. 3, Nos. 5/6, pp. 129–135.) Discussion of the connection between variations of the temperature at the tropopause and in the upper troposphere and lower stratosphere.

551.510.52 : 546.214 : 551.524.7 625

**Total Ozone Related to Troposphere Temperatures.**—(*Tech. Bull. nat. Bur. Stand.*, Nov. 1949, Vol. 33, No. 11, pp. 128–129.) Long-term measurements of the ultraviolet radiation reaching the earth from the sun reveal a correlation between the temperature of the air at an altitude of about 5 miles and the total amount of $O_3$ in the atmosphere.

551.510.535 626

**Ionization of the Stratosphere over High- and Low-Pressure Regions of the Troposphere.**—G. Falkenberg & E. Lauter. (*Z. Met.*, May/June 1949, Vol. 3, Nos. 5/6, pp. 136–140.) A statistical analysis of the results of 650 observations of the reflection of 1 250-m waves from the E layer.

551.510.535 627

**Calculation of the Absorption Decrement for a Parabolic Ionospheric Layer for the case of Vertical Incidence.**—E. Argence & K. Rawer. (*C. R. Acad. Sci., Paris*, 14th Nov. 1949, Vol. 229, No. 20, pp. 996–997.) The total absorption is the sum of three parts, the first ($\delta_1$) being the attenuation which depends on the path length, the second ($\delta_2$) the loss of energy due to collisions in the layer, and the third ($\delta_3$) due to partial reflection by the sporadic-E layer. Formulae are obtained for $\delta_2$ for the case of total reflection and for that of layer penetration ; the influence of the earth's magnetic field is neglected. These formulae are of importance in the study of E-layer absorption, which cannot be neglected when working near the critical frequency. See also 3035 and 3879 of 1949.

551.510.535 628

**Electron Production in the Ionosphere.**—R. Seeliger. (*Naturwissenschaften*, Nov. 1949, Vol. 36, No. 11, pp. 321–327. Bibliography, p. 327.) A study of conditions in the E layer.

551.510.535 629

**True Heights of the $F_2$ Layer.**—G. R. White & I. S. Wachtel. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 239–242.) " The true heights of electron density for Washington were computed, using the parabolic method, for winter, summer, and equinox of 1945 at noon and at midnight. Representative curves were formed by combining the data from a large number of days." The distributions obtained show that the maximum electron-density of the $F_2$ region occurs at a lower height at midday than at midnight. The curves are compared with similar curves for Huancayo, where the height of maximum electron density is greater at midday than at midnight.

551.510.535 630

**Ionospheric Research in the U.S. Air Force.**—N. C. Gerson. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 286–287.) Summary of Pennsylvania State College Conference paper. Various experimental undertakings planned or in progress are briefly mentioned.

551.510.535 631

**Ionospheric Research at the Pennsylvania State College.**—B. B. Underhill. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 287–288.) Summary of Pennsylvania State College Conference paper. An account of the history, organization, and research in progress at the Radio Propagation Laboratory.

551.510.535 632

**Summary of a Meeting on Mathematical Problems in Ionospheric Research.**—N. C. Gerson. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 289–290.) Discussion following the Pennsylvania State College Conference. Ten of the major problems suggested for investigation are noted.

551.510.535 633

**Sunrise Effects in F Region from High Speed Ionospheric Recordings.**—H. W. Wells. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 277–280.)

551.510.535 : 525.624 634

**Tidal Effects in the F Layer.**—M. W. Jones & J. G. Jones. (*Phys. Rev.*, 15th Aug. 1949, Vol. 76, No. 4, p. 581.) A periodic variation of the thickness of the F layer, noted at College, Alaska, was submitted to harmonic analysis. In addition to the diurnal change due to the sun's ionizing radiation, variations resulting from the contractions and expansions of the layer due to solar and lunar gravitational forces were noted. The lunar diurnal wave, whose amplitude varied from 0.1 to 1.6 times that of the solar semidiurnal tide during the winter of 1948–1949, can be explained by simple tidal theory.

551.510.535 : 525.624 635

**Lunar Ionospheric Variations at Low-Latitude Stations.**—A. G. McNish & T. N. Gautier. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 303–304.) Amplitudes and phase angles for variation in $F_2$-layer critical frequency at noon and 4 p.m. are tabulated for Huancayo, Christmas Island and Leyte (near the geomagnetic equator) and Trinidad and Maui (about 20° from the geomagnetic equator). Results are given for solstices and equinoxes; the age of the moon is included. The phase reversal in the noon values at 20° with respect to 0° geomagnetic latitude may be associated with the equatorial trough in the latitude distribution of $F_2$-layer critical frequencies.

551.510.535 : 525.624 636

**On the Practical Determination of Lunar and Luni-Solar Daily Variations in Certain Geophysical Data.**—K. K. Tschu. (*Aust. J. sci. Res., Ser. A*, March 1949, Vol. 2, No. 1, pp. 1–24.)

551.510.535 : 621.396.11 637

**Summary of Symposia** [at the Pennsylvania State College Conference].—A. H. Waynick. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 290–294.) The symposia covered (*a*) long-wave propagation, (*b*) the physical bases for the Appleton-Hartree dispersion equation, (*c*) studies of the auroral region, (*d*) ionospheric absorption, theory and measurement, (*e*) tidal oscillations in the upper atmosphere.

551.510.535 : 621.396.11 638

**The Effect of the Lorentz Polarization Term on the Vertical Incidence Absorption in a Deviating Ionosphere Layer.**—J. M. Kelso. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, p. 284.) Summary of Pennsylvania State College Conference paper. Appropriate assumptions and approximations are used to derive expressions for absorption and for group height by means of the Lorentz theory. The results are compared with those obtained by Hacke from the Sellmeyer theory. The determination of the value of the collisional frequency at the level of maximum ionization is discussed for both theories. See also 1050 of 1949 (Hacke & Kelso).

551.510.535 : 621.396.11 639

**Sporadic Ionization at High Latitudes.**—Meek. (*See* 725.)

551.524 + 551.54 640

**Atmospheric Pressure and Temperature Gradients over Central Europe.**—R. Höhn. (*Z. Met.*, May/June 1949, Vol. 3, Nos. 5/6, pp. 148–153.)

551.594.5 : 621.396.11 641

**Sky-Wave Observations ($\lambda = 1250$ m) during Displays of the Northern Lights in 1947.**—Lauter & Sprenger. (*See* 728.)

## LOCATION AND AIDS TO NAVIGATION

621.396.9 642

**The Use of Radar in the Ice-Breaker Service.**—H. Larsson. (*J. Inst. Nav.*, Oct. 1949, Vol. 2, No. 4, pp. 315–323.) A R.C.A. Type CR-101 radar set was used for the 3-cm band and a Raytheon Mariners' Path-finder for the 10-cm band. The type of echo received from various types of ice is described. Radar can be used to determine the ice-breaker's position from bearings on landmarks, to detect open-water channels in conditions of bad visibility, or to locate a ship whose position is only approximately known in the ice-breaker. The relative merits of the two wavelengths are discussed.

621.396.9 : 523.53 643

**A Study of Radio Reflections from Meteor Trails in Research on the Upper Atmosphere.**—McNish. (*See* 613.)

621.396.9 : 551.515.2 644

**Industrial Radar for Hurricane Tracking.**—W. F. Gerdes & R. C. Jorgensen. (*Science*, 7th Oct. 1949, Vol. 110, No. 2858, pp. 357–360.) Hurricanes present a distinctive picture on a radar set because of the accompanying heavy rain. Modifications to a 10-cm radar set Type SCR-784, for hurricane observation near Freeport, Texas, are discussed.

621.396.933 645

**High-Stability Radio Distance-Measuring Equipment for Aerial Navigation.**—H. Busignies. (*Proc. Instn Radio Engrs, Aust.*, Nov. 1948, Vol. 9, No. 11, pp. 5–10.) Paper presented at the 1948 Australian I.R.E. conference, Sydney, and describing equipment noted in 1402 and 1944 of 1949.

621.396.933 646

**Improved Radio Systems for Modern Aircraft.**—(*Tele-Tech*, Nov. 1949, Vol. 8, No. 11, pp. 31–33, 59.) Discussion of recent developments incorporating many P.I.C.A.O. recommendations (see 2436 and 3532 of 1947) and of distance measuring equipment, full use of which may be expected by 1953. See also 102 of January (Burgmann).

621.396.933 647

**The Oboe System.**—P. Besson. (*Onde élect.*, Oct. & Nov. 1949, Vol. 29, Nos. 271 & 272, pp. 351–367 & 414–426.) A comprehensive description. See also 1790 of 1947 (Jones).

621.396.933 648

**Pulse Navigation Systems.**—W. L. Barrow. (*Proc. Instn Radio Engrs, Aust.*, Oct. 1949, Vol. 10, No. 10, pp. 276–282.) Paper presented at the International Meeting on Marine Radio Aids to Navigation, New York, 1947. Brief review of short-, medium- and long-distance systems available at present, with discussion of the factors affecting their suitability for particular services.

621.396.933 649

**Rotating-Field Radio Beacons for Navigation.**—S. Ostrovidow. (*Rev. gén. Élect.*, Nov. 1949, Vol. 58, No. 11, pp. 469–473.) The general theory for this type of beacon is outlined and an installation near Marseilles is described. This operates on 3.965 Mc/s with modula-

tion frequency of 1 kc/s. Aerial power is about 2 kW and the effective range above 500 km. Sources of error, in particular those due to the site, are discussed. Performance is considered to be comparable with that of the Decca and Gee systems.

621.396.933(083.71) 650

**Standards on Radio Aids to Navigation: Definitions of Terms, 1949.**—(*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1364–1371.) Prepared by an I.R.E. committee.

621.396.9 : 621.396.67 651

**Radar Scanners and Radomes.** [Book Review]—W. M. Cady, M. B. Karelitz & L. A. Turner (Eds). Publishers: McGraw-Hill, New York and London, 1949. 491 pp., 42s. (*Nature, Lond.*, 24th Dec. 1949, Vol. 164, No. 4182, pp. 1064–1065.) "No. 26 of the M.I.T. Radiation Laboratory series. The book "deals mainly with the engineering aspects of the design of scanners and their housings. . . . There is almost no reference to equipment operating at wavelengths greater than 10 cm, and there is little discussion of scanners not developed at the Radiation Laboratory."

## MATERIALS AND SUBSIDIARY TECHNIQUES

533.56 652

**A New Type of Diffusion Pump.**—E. L. Harrington. (*Rev. sci. Instrums.*, Nov. 1949, Vol. 20, No. 11, pp. 761–762.) Improvements to the Kerth Hg diffusion pump, reducing noise, operating expenses and operating temperature, and providing a qualitative measure of the degree of vacuum. The introduction of a Hg seal between the high- and low-vacuum sections makes interrupted operation possible.

535.37 653

**Luminescence of Barium-Strontium Oxide.**—R. E. Aitchison. (*Nature, Lond.*, 24th Dec. 1949, Vol. 164, No. 4182, p. 1088.) Well-activated (Ba,Sr)O cathodes gave a pale blue fluorescence under electron bombardment, the total emission spectrum extending from 4 600 Å to 5 400 Å. The intensity of the emitted radiation decreases with increasing oxide temperature and falls rapidly above 700° K. Afterglow is negligible.

535.37 654

**The Luminescence Characteristics of Tin-Activated Zinc Sulphide Phosphors.**—G. F. J. Garlick & D. E. Mason. (*Proc. phys. Soc.*, 1st Dec. 1949, Vol. 62, No. 360A, pp. 817–822.) "It has been found that the inclusion of stannous compounds in relatively large concentrations in the preparation of zinc sulphide phosphors results in an intense red luminescence when excitation is by ultra-violet light of long wavelength (3 650–4 000 Å). Such characteristics as the luminescence spectra, excitation spectra, phosphorescence and thermoluminescence and the variation of luminescence with temperature have been studied. The hexagonal crystal form of zinc sulphide is essential to the production of efficient phosphors with tin as the activating impurity."

537.228.1 : 548.0](083.7) 655

**Standards on Piezoelectric Crystals, 1949.**—(*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1378–1395.) Prepared by an I.R.E. committee. Includes (*a*) definitions of axes for the various crystal systems, (*b*) specification of crystal-plate orientation, (*c*) basic equations, symbols and units of piezoelectric theory.

538.221 656

**On the Conditions for the Occurrence of Ferromagnetism in Metal Compounds.**—J. H. Gisolf. (*Physica, 's Grav.*, Sept. 1949, Vol. 15, Nos. 8/9, pp. 677–678. In English.) It is necessary, though not sufficient, that the lines of communication between the metal atoms which are in exchange interaction form a 3-dimensional network. The ferrites with spinel structure are considered; it appears that only the metal ions in tetrahedron holes can contribute to the ferromagnetism.

538.221 657

**Relation between the Thermal Expansion, the Curie Temperature and the Lattice Spacing of Homogeneous Ternary Nickel-Iron Alloys.**—J. J. Went. (*Physica, 's Grav.*, Sept. 1949, Vol. 15, Nos. 8/9, pp. 703–710. In English.) These quantities were measured for 15 different alloys, all having about 50% Ni and 50% Fe. The change in the expansion anomaly is closely related to the change in Curie temperature. The value of the latter change depends upon the relative positions of Ni and the third alloying element in the periodic system of elements. There is no direct relation between the change in Curie temperature and the lattice spacing.

538.221 658

**Ferromagnetic Alloys in the Systems Cu-Mn-In and Cu-Mn-Ga.**—F. A. Hames & D. S. Eppelsheimer. (*J. Metals*, formerly *Metals Technol.*, Aug. 1949, Vol. 1, No. 8, pp. 495–499.) An investigation to determine whether ferromagnetic $\beta$-phases exist in these systems analogous to such phases in the systems Cu-Mn-Al and Cu-Mn-Sn. A Cu-Mn-In alloy having an ordered body-centred cubic structure analogous to that of the Heusler alloys is discussed. An alloy $Cu_{1.87}Mn_{1.16}Ga_{1.98}$ variously heat treated is feebly magnetic and has a 2-phase structure. With higher Cu contents, quenched Cu-Mn-Ga alloys are strongly magnetic.

538.221 659

**Magnetic Ferrites.**—C. L. Snyder, E. Albers-Schoenberg & H. A. Goldsmith. (*Elect. Mfg, N.Y.*, Dec. 1949, Vol. 44, No. 6, pp. 86–91.) "Ferramics" consist only of metallic oxides, and have high resistivity and high permeability, but low losses. They can thus be used to reduce the size and weight of h.f. magnetic apparatus. Properties of typical ferramics are tabulated or shown graphically. See also 3447 of 1948 (Snoek).

538.221 660

**Magnetic Ferrites for High-Frequency Uses.**—F. G. Brockman. (*Elect. Engng, N.Y.*, Dec. 1949, Vol. 68, No. 12, pp. 1077–1080.) Discussion of (*a*) the historical background of research on such materials by the Philips Co. at Eindhoven, Holland, and Irvington-on-Hudson, N.Y., U.S.A., with special attention to the crystallography of the ferrites, (*b*) characteristics of the materials, (*c*) applications, (*d*) the existence of displacement currents as well as ohmic currents in a core of ferroxcube III. See also 2265 of 1948 (Verwey, Haayman & Romeyn) and 2387 of 1948 (Rinia et al.).

538.221 : 621.775.7 661

**The Coercive Field of Ferromagnetic Powders.**—C. Guillaud. (*C. R. Acad. Sci., Paris*, 24th Oct. 1949, Vol. 229, No. 17, pp. 818–819.) The variation of coercive field with temperature for a powder consisting of single crystals of $Mn_2Sb$ is discussed.

538.23 : 621.318.323.2 662

**Unsymmetrical Hysteresis Loops in a Nickel-Iron Alloy.**—J. L. Rothery & An Wang. (*Nature, Lond.*, 10th Dec. 1949, Vol. 164, No. 4180, pp. 1004–1005.) A very pronounced asymmetry of the hysteresis loop has been observed in one particular sample of grain-oriented Ni-Fe alloy core when a large d.c. magnetizing pulse was applied to the core and followed by smaller a.c. excitation. If the a.c. excitation is gradually increased until saturation is reached and then reduced, the asymmetry persists except during saturation. Reversal of the

magnetizing pulse reverses the loop. The phenomenon persists even if the core is completely disconnected for several days or weeks; the core can therefore be used to store information about the polarity of the last transient magnetizing pulse.

546.431.82 663

**Theory of Barium Titanate: Part 1.**—A. F. Devonshire. (*Phil. Mag.*, Oct. 1949, Vol. 40, No. 309, pp. 1040–1063.) Dielectric and crystallographic properties are considered. By expanding the free energy as a function of polarization and strain, and making reasonable assumptions about the coefficients, the various crystal transitions can be explained. The dielectric constants, crystal strains, internal energy and self polarization are calculated as functions of temperature. Relations are obtained between the coefficients in the free energy and the ionic force constants. These are used to estimate some of the coefficients which are not completely determined by experimental data.

620.197.7 664

**Preventing Fungus Damage.**—E. F. Little. (*Elect. Mfg, N.Y.*, Nov. 1949, Vol. 44, No. 5, pp. 89–91, 170.) Fungus attack is to be expected in most climates, but particularly under hot, moist conditions. Materials most liable to attack are those containing organic compounds. Protection is secured by the addition of fungicides harmless to animal life, notably copper-8-quinolinolate and copper naphthanate.

621.315.5/.6 665

**Plastics can be Electrical Conductors.**—(*Elect. Mfg, N.Y.*, Nov. 1949, Vol. 44, No. 5, pp. 60–63 . . 180.) Discussion of Markite materials. For another account see 3443 of 1949.

621.315.59 666

**Conductivity in Semiconductors.**—K. Lark-Horovitz. (*Elect. Engng, N.Y.*, Dec. 1949, Vol. 68, No. 12, pp. 1047–1056.) A survey of papers presented at the A.I.E.E. summer general meeting.

621.315.59 667

**On the Spontaneous Current Fluctuations in Semiconductors.**—J. H. Gisolf. (*Physica, 's Grav.*, Sept. 1949, Vol. 15, Nos. 8/9, pp. 825–832. In English.) Formulae are derived which indicate that (*a*) the effective current strength of the fluctuations is proportional to the field strength, (*b*) at l.f. the fluctuations are independent of frequency, (*c*) at h.f. the effective current strength decreases with increasing frequency, (*d*) for alternating fields, the fluctuations depend on the difference between the observation frequency and that of the alternating field, (*e*) effective current strength is proportional to the square root of the mean life time of the free electrons in the semiconductor.

621.315.59 : 535.215.1 668

**Photoeffects in Semiconductors.**—(*Elect. Engng, N.Y.*, Nov. 1949, Vol. 68, No. 11, pp. 937–942.) Long summary, compiled by J. A. Becker, of 3 papers read at an A.I.E.E. symposium on 'Electrical Properties of Semiconductors and the Transistor', namely:—General Features of Photoconductivity and Photoemission in Semiconductors, by L. Smith. External Photoelectric Effects in Semiconductors, by L. Apker. Internal Photoeffects in Germanium, by J. N. Shive. For summaries of other papers at this symposium, see 240 of January.

621.318.22 669

**Magnetic Saturation Intensities and Curie Temperatures for Some Industrial Permanent Magnet Materials.**—L. Ward. (*Metallurgia, Manchr*, Nov. 1949, Vol. 41, No. 241, pp. 3–7.) Magnetic analysis of various steels and alloys indicates that in alloys of the Alnico type there is a reversible phase up to the Curie point, and that cobalt steels, as industrially heat treated, contain mixtures of austenitic and magnetic-α phases. The apparatus used and the experimental procedure are described.

621.318.22 670

**Mechanism of the Coercive Field and of the Remanent Magnetism of Powdered MnBi. Generalization.**—C. Guillaud. (*C. R. Acad. Sci., Paris*, 14th Nov. 1949, Vol. 229, No. 20, pp. 992–993.) The coercive field of ordinary permanent magnets is due to internal strains. Fields obtained with powdered MnBi are of a much higher order of magnitude. It is supposed that in the absence of an inductive field, the elementary magnetic vectors are all parallel, but half of one sense and half of the opposite sense. A sufficient inductive field causes one set of these vectors to rotate so that all are thereafter of the same sense. The effect is more pronounced the smaller the grain size; the cases where one grain includes (*a*) several elementary Weiss domains, (*b*) at most one domain, are considered. This mechanism also exists in certain ferromagnetic substances which belong to the hexagonal system, and also for substances belonging to the cubic system. To obtain the best characteristics in magnets produced from powdered materials, each grain should contain only a single elementary domain and the axes of easy magnetization should be parallel.

669.157.82 : 621.775.7 : 537.311.3 671

**Electrical Resistivity Measurements on Iron-Silicon Compacts prepared by the Powder Metallurgy Procedure.**—F. W. Glaser. (*J. Metals*, formerly *Metals Technol.*, Aug. 1949, Vol. 1, No. 8, pp. 475–480.) A study, by means of electrical resistivity measurements, of the progress of diffusion in Fe-Si alloys.

## MATHEMATICS

681.142 672

**The Binary Quantizer.**—K. H. Barney. (*Elect. Engng, N.Y.*, Nov. 1949, Vol. 68, No. 11, pp. 962–967.) A detailed description is given of apparatus for the continuous conversion of input currents or voltages, varying with time, into discrete binary numbers. When a difference exists between the input voltage and the feedback voltage from the counter circuit, the direction of operation of the counter is controlled by the sign of this difference or error voltage.

681.142 673

**An Analogue Computer for the Solution of Linear Simultaneous Equations.**—R. M. Walker. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1467–1473.)

681.142 : 517.512.2 674

**Fourier Coefficient Harmonic Analyzer.**—S. Charp. (*Elect. Engng, N.Y.*, Dec. 1949, Vol. 68, No. 12, p. 1057.) Summary only. A ball-and-roller type of mechanical integrator, specially adapted for determining Fourier coefficients directly. The curve to be analysed is drawn on graph paper, which is wrapped round a cylinder and inserted into the machine. The curve is tracked manually; counters then give the required coefficients. Each time the curve is scanned another set of coefficients is determined. A range of 100 harmonics can be investigated. The time required for each harmonic is 2–7 minutes, depending on the curve.

681.142 : 621.318.572 675

**Gate-Type Shifting Register.**—J. H. Knapton & L. D. Stevens. (*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 186..192.) Double-triode trigger circuits are used to store the numbers 1 or 0 of the binary system, and each stage passes its number to a following stage on receipt of a shift pulse. The register operates reliably with 1-μs pulses at 250 kc/s and numbers have been circulated through six stages at 600 kc/s.

## MEASUREMENTS AND TEST GEAR

621.317.2 : 620.193 676

**A Laboratory for the Study of the Resistance of Materials under Tropical Conditions.**—A. Nizery & S. Crespi. (*Rev. gén. Élect.*, Nov. 1949, Vol. 58, No. 11, pp. 455–468.) A full description of the laboratory equipment at Saint-Cyr is preceded by a study of the actual conditions to be reproduced.

621.317.3 : 621.396.822 677

**The Application of I.F. Noise Sources to the Measurement of Over-All Noise Figure and Conversion Loss in Microwave Receivers.**—L. A. Moxon. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1433–1437.) A method is described of extending noise-diode technique to microwaves, using a frequency changer to produce the required r.f. test signal from an i.f. noise source. An experimental procedure has been developed which enables the signal level to be accurately evaluated, subject to such limitations as those of Dicke's reciprocity theorem, which are of little practical significance for the applications so far considered. As in the low-frequency case, measurements are in general independent of receiver bandwidth, there is no stray radiation problem, and the receiver output indicating device may follow any law within wide limits. The necessary components are easy to construct and can be calibrated without the use of additional apparatus. The system is particularly well suited to mixer crystal measurements, since both conversion loss and over-all noise figure can be measured with equal facility, and the latter is readily analysed in terms of conversion loss and noise ratio. See also 1196 of 1947.

621.317.3(083.74) : 621.396.931 678

**Standards on Railroad and Vehicular Communications: Methods of Testing, 1949.**—(*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1372–1375.) Prepared by an I.R.E. committee.

621.317.324† : 621.396.11 679

**The Fresnel Reflection Formula in the Region of Long Waves.**—E. Roeschen. (*Funk u. Ton*, 1949, Vol. 3, Nos. 9/10, pp. 525–528.) A short description of a method of verification by determining the field distribution beneath a vertical radiator. A 30-m Al mast was mounted on a platform floating on a lake and an earth plate of area 4 m² was arranged symmetrically below the mast. The field distribution at different levels above and below the plate was measured for wavelengths of 217, 437 and 1450 m. The results are shown graphically.

621.317.335.2† : 621.319.4(083.74) 680

**Standard Test Capacitors.**—(*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 168–174.) A T-network of ordinary mica capacitors can provide a very small capacitance (e.g. 0.001 pF) between the extremities of the series elements when the shunt arm is grounded, if the capacitors in the series arms are of only a few picofarads each and the capacitance in the shunt arm is several thousand picofarads. Such networks can be used as low capacitance standards in the measurement of interelectrode capacitances.

621.317.35 : 621.3.015.33 681

**Analysis of Pulse Shape.**—G. F. Draffin. (*Proc. Instn Radio Engrs, Aust.*, Nov. 1948, Vol. 9, No. 11, pp. 10–15.) A simple graphical method is described for determining the amplitudes of the various harmonics present in a pulse of any waveform. The method is particularly useful for the rapid study of such problems as the effect on pulse shape of frequency-dependent amplification or attenuation.

621.317.373.029.64 682

**The Experimental Determination of Phase Characteristics of Circuits for Centimetre Waves.**—M. Denis & P. Pallud. (*Ann. Radioélect.*, Oct. 1949, Vol. 4, No. 18, pp. 315–330.) A critical review of the principles, methods and apparatus used for quantitative measurements of (*a*) phase displacement between input and output at a fixed frequency; (*b*) phase variations with frequency, and phase distortion; (*c*) phase velocity in transmission lines; (*d*) phase variation of active quadripoles as a function of certain parameters. The relative advantages and disadvantages of dynamic, reflection, resonance and transmission methods of measurement are discussed. Illustrations and some details are given of suitable wavemeters and slotted lines.

621.317.43 : 669.14-41 683

**Core-Loss Test for Narrow Silicon-Steel Strip.**—J. A. Ashworth. (*Bell Syst. tech. Publ. Monogr.*, B-1667, 8 pp.) Description of a method for measuring the across-grain core loss in silicon-steel strip when the standard A.S.T.M. method cannot be applied. The Epstein method is modified by decreasing the lengths of the test windings and test pieces.

621.317.66 684

**Measurement of Microwave-Transmission Efficiency.**—A. L. Cullen. (*Radio tech. Dig., Édn franç.*, Dec. 1949, Vol. 3, No. 6, pp. 345–348.) French version of 3196 of 1949.

621.317.7 685

**A New Type of Electrical Instrument.**—K. A. Pullen. (*Instruments*, Nov. 1948, Vol. 21, No. 11, pp. 1008–1011.) A double-moving-coil arrangement is used, to overcome difficulties caused by vibration, shock etc. The coils are mounted on opposite ends of the support frames, which carry the pivots at the centre. The pointer is an extension of the upper frame. Theory and mechanical details of such instruments are fully discussed. Advantages include high torque/weight ratio, high sensitivity, low power consumption and extreme robustness.

621.317.7.089.6 686

**Equipment for Instrument Calibration.**—E. A. Gilbert. (*Elect. Engng, N.Y.*, Dec. 1949, Vol. 68, No. 12, pp. 1065–1066.) Description of three self-contained units for the calibration of all types of a.c. and d.c. measuring instrument now used, and designed to anticipate future requirements. A power supply at 110–120 V with frequency between 50 and 1600 c/s is required.

621.317.725 687

**A New Expanded-Scale A.C. Voltmeter.**—N. P. Millar. (*Elect. Engng, N.Y.*, Dec. 1949, Vol. 68, No. 12, p. 1044.) Summary only. An iron-core saturable reactor is used to energize the moving coil of a standard electrodynamic wattmeter. Accuracy is within about ½% of full-scale rating.

621.317.73.029.63 688

**Circular Standing-Wave Detector.**—(*Radio & Televis. News, Radio-Electronic Engng Supplement*, Dec. 1949, Vol. 13, No. 6, pp. 12–13.) A brief general description of the apparatus. For another account see 3198 and 3203 of 1949 (Meinke).

621.317.733 689

**Self-Balancing Resistance Bridge.**—H. F. Rondeau. (*Gen. elect. Rev.*, Oct. 1949, Vol. 52, No. 10, pp. 45–46.) A servo system, consisting of an electronic converter and amplifier unit feeding a reversible induction motor, balances the bridge rapidly and automatically. The equipment, which can be switched for manual operation, is recommended for production tests requiring an accuracy of about 1% over the resistance range of 0.01Ω to above 30 kΩ.

621.317.74 : 621.315.212 690

**Maintenance Equipment for Coaxial-Cable Circuits.**—J. Selz. (*Câbles & Transmission, Paris*, Oct. 1949, Vol. 3, No. 4, pp. 306–325.) Specifications and description of apparatus constructed in France for making comprehensive line measurements in the range 30 kc/s–4 Mc/s.

621.317.762 691

**Loading and Coupling Effects of Standing-Wave Detectors.**—K. Tomiyasu. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1405–1409.) "When measuring impedances on transmission lines, insensitive standing-wave detectors have the effect of yielding lower standing-wave ratios than the true values. Double-hump distribution curves are shown to be the result of very tight coupling of the detector. Detectors that can be represented by a susceptance component may indicate unsymmetrical distribution curves. Sensitive detectors used on transmission lines having low power levels can introduce tight-coupling effects. Conditions are given for a loosely coupled detector.

621.317.79 : 621.396.615.12 692

**Low Frequency Generators.**—R. A. Raffin-Roanne. (*Radio prof., Paris*, Nov. 1949, Vol. 18, No. 178, pp. 16–20.) Description of two signal generators for testing l.f. amplifier fidelity. The first provides a sinusoidal output of 500 mW maximum with <3% harmonic distortion, covering 15 to 16 000 c/s. The second is an adapter for use with the first to provide a rectangular-wave output. Complete circuit diagrams are given and some aspects of amplifier fidelity are considered.

621.396.933.001.6 693

**The Servicing of Airborne Equipment.**—T. R. W. Bushby. (*Proc. Instn Radio Engrs, Aust.*, July 1949, Vol. 10, No. 7, pp. 190–195.) 1948 Australian I.R.E. Convention paper. An outline of the organization of servicing procedure in Australia is given. Servicing requirements and test apparatus are discussed. Instruments for checking transmitter radiation and ignition noise and for the overhaul of relays, microphones and headphones are described.

## OTHER APPLICATIONS OF RADIO AND ELECTRONICS

526.9 : 621.396.9 694

**Geodetic Measurements by Radar.**—D. J. Halliday. (*Nature, Lond.*, 10th Dec. 1949, Vol. 164, No. 4180, pp. 1005–1006.) An aircraft flies in a circular arc at the centre of which is a ground radar beacon. The distance of the aircraft from a second beacon is recorded at regular intervals. The geodetic distance between the beacons is calculated from the radius of the arc and the minimum recorded distance from the second beacon, corrections being made for altitude and atmospheric refraction.

531.78 : 621.3.083.7 695

**The Differential Transformer as Applied to Instrumentation.**—W. D. Macgeorge. (*Science*, 7th Oct. 1949, Vol. 110, No. 2858, pp. 365–368.) Basic design of differential transformers is discussed and illustrated by a description and performance details of the 'Atcotran' type made by the Automatic Temperature Control Co. Suitable circuits for using these transformers in the measurement of variables such as pressure, flow and force are also considered, the null-balance type of circuit being especially noted. Practical wiring procedures are given for systems in which the indicating instrument is a considerable distance from the equipment on which measurements are to be made.

535.61-15 696

**Some Experimental Demonstrations with Infra-Red Rays.**—T. Gast. (*Funk u. Ton*, 1949, Vol. 3, Nos. 9/10, pp. 529–533.)

538.71 697

**Airborne Magnetometer for Measuring the Earth's Magnetic Vector.**—E. O. Schonstedt & H. R. Irons. (*Science*, 7th Oct. 1949, Vol. 110, No. 2858, pp. 377–378.) The airborne total field magnetometer, developed during World War II by the Naval Ordnance Laboratory and Bell Telephone Laboratories, has been modified so that the total magnetic field vector is measured instead of the intensity only.

539.16.08 698

**Equipment for the Counting of Particles by the Geiger-Müller Method.**—A. Berthelot. (*J. Phys. Radium*, July 1946, Vol. 7, No. 7, pp. 185–192.) Description of apparatus for laboratory measurements, including scaler circuits, a stabilized h.v. supply, pulse generator, and a complete unit comprising preamplifier, decade scaler circuit, power stage, counter, power supply and voltmeter.

539.16.08 699

**Glass-Walled Counters with External Cathode.**—R. Maze. (*J. Phys. Radium*, June 1946, Vol. 7, No. 6, pp. 164–166.) Colloidal graphite is deposited on the outside of a thin-walled tube of ordinary glass with axial wire anode and hydrogen filling. In use as a counter, the inside of the glass becomes charged as uniformly as if it had a conducting surface layer. The arrangement is thus equivalent to the usual type of counter with the cathode earthed through a high resistance in parallel with a capacitance. Owing to the weak secondary effects of glass a production efficiency of 100% can easily be obtained.

539.16.08 700

**Ion Mobilities in Geiger-Müller Counters.**—H. den Hartog & F. A. Muller. (*Physica, 's Grav.*, Sept. 1949, Vol. 15, Nos. 8/9, pp. 789–800. In English.) Formulae and measuring instruments are described for a detailed study of the deionization current in a self-quenching counter. Ion mobility in alcohol is 2.6 cm/sec per V/cm, in argon 7.1 cm/sec per V/cm, both at 100 mm Hg pressure. These results supersede earlier measurements by van Gemert and the authors, noted in 1139 of 1949.

539.16.08 701

**Spurious Counts in Geiger Counters and the Pre-treatment of the Electrodes.**—J. D. Louw & S. M. Naudé. (*Phys. Rev.*, 15th Aug. 1949, Vol. 76, No. 4, pp. 571–572.)

539.16.08 702

**On the Temperature Dependence of Counter Characteristics in Self-Quenching G-M Counters.**—O. Parkash. (*Phys. Rev.*, 15th Aug. 1949, Vol. 76, No. 4, pp. 568–569.)

551.508.1 : 621.317.083.7 703

**Range-Adjusting Radiosonde Recorder.**—G. E. Beggs, Jr. (*Elect. Engng, N.Y.*, Nov. 1949, Vol. 68, No. 11, p. 990.) Summary of A.I.E.E. paper. Range-adjustment and calibration are performed automatically. A block diagram of the complete system is included.

621.316.718 704

**Basic Control Requirements of D.C. Adjustable-Voltage Drives.**—E. E. Moyer & M. E. Cummings. (*Elect. Mfg, N.Y.*, Nov. 1949, Vol. 44, No. 5, pp. 64–70.) Discussion of the field control sequences for starting, accelerating, running, retarding, reversing and stopping the motor unit of a d.c. adjustable-voltage drive of the Ward-Leonard type. The necessary design information is given mainly in a comprehensive series of scaled diagrams, with brief notes.

621.316.726 : [621.365.55† + 615.84 705

**Industrial Oscillator Frequency Control.**—J. W. Lower. (*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 84–86.) Describes an a.f.c. system for diathermy or dielectric-heating power oscillators. The oscillator frequency is

sampled 200 times per sec and compared with a crystal-controlled reference frequency. The resulting error signal is applied to a servo system operating a 2-phase induction motor which adjusts the reactive tuning element of the oscillator, whose frequency is thus kept within 0.01 % of the 27-Mc/s standard.

621.317.39 : 534.08 706

**A New Method of Vibration Measurement for the Frequency Range 20 to 20 000 c/s.**—F. Massa. (*Instruments*, Nov. 1948, Vol. 21, No. 11, pp. 1012–1014.) An accelerometer unit is designed so that its electromechanical system has a single degree of freedom and its resonance frequency is above 20 kc/s. The unit consists of a system of ADP crystal plates mounted within a cylindrical stainless steel housing 1 in. in diameter and 2 in. long. A stress is imparted along the piezoelectric axis of the plates in proportion to the component of mechanical vibration in the direction of the cylinder axis. Typical response curves are shown and discussed.

621.365.54† 707

**Production Economies realized by Proper Use of Induction Heating.**—J. A. Evans. (*Materials & Methods*, Nov. 1949, Vol. 30, No. 5, pp. 57–60.)

621.38.001.8 708

**Electronics Symposium, 1949.**—(*Engineer, Lond.*, 4th Nov. 1949, Vol. 188, No. 4893, p. 525.) Brief notice of the papers presented and the apparatus exhibited at a symposium on instruments in research and industry, held in London 2nd–4th Nov. 1949.

621.38.001.8 709

**Electronic Aids to Industry.**—(*J. Brit. Instn Radio Engrs*, Dec. 1949, Vol. 9, No. 12, pp. 446–464.) A survey prepared by the Technical Committee of the British I.R.E. Applications are grouped in 33 sections, each with a bibliography which, though not exhaustive, may serve as a guide to more detailed technical information.

621.38.001.8 : 543/545 710

**Electronic Instrumentation for Chemical Laboratories.**—F. Gutmann. (*Proc. Instn Radio Engrs, Aust.*, Sept. 1949, Vol. 10, No. 9, pp. 241–254. Bibliography, pp. 254–256.) Paper presented at the Australian I.R.E. convention, Sydney, 1948. A comprehensive survey, with many diagrams of the principles of various analysing instruments, including pH meters and the associated current amplifiers, colorimeters using photoelectric cells, polarographs for the analysis of electrolytes by measurement of the current/voltage curve, gas analysers, titration potentiometers, electrolytic conductance bridges, dielectric-constant and dielectric-loss measuring equipment for radio frequencies, moisture meters, supersonic equipment, an ionization gauge and an electronic balance.

## PROPAGATION OF WAVES

001.3 : 551.510.535 : 538.566 711

**The Conference on Ionospheric Research.**—(*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 281–294.) Summaries of symposia and abstracts of papers presented at the conference held at Pennsylvania State College under joint sponsorship of that College and of the Geophysical Research Directorate, U.S.A.F., 27th–29th June 1949. Selected papers are noted below in this section, or in the 'Geophysical and Extraterrestrial Phenomena' section.

538.56 712

**General Expression of Huyghens' Principle for Electromagnetic Waves in an Imperfectly Transparent Medium.**—J. Brodin. (*C. R. Acad. Sci., Paris*, 21st Nov. 1949, Vol. 229, No. 21, pp. 1064–1066.) Extension of 523 above to the case of e.m. waves. The layers involved in this case consist respectively of electric and magnetic dipoles.

538.566 713

**On the Electromagnetic Surface Wave of Sommerfeld.**—T. Kahan & G. Eckart. (*Phys. Rev.*, 1st Aug. 1949, Vol. 76, No. 3, pp. 406–410.) See 2892 of 1949.

538.566.3 : 551.510.535 714

**Application of the Magneto-Ionic Theory to Radio Waves Incident Obliquely upon a Horizontally-Stratified Ionosphere.**—H. G. Booker. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 243–274.) Pennsylvania State College Conference paper. "A study is made of the propagation of wave-packets incident obliquely upon a slowly varying plane-stratified ionosphere, allowing for the full effect of the earth's magnetic field. Propagation of a magneto-ionic component is described in terms of the vertical component $q$ of the phase-propagation vector, so as to avoid using a refractive index which depends in a complicated way on an unknown angle of refraction. The fundamental formula of the theory is an algebraic quartic equation for $q$, the coefficients in which depend on electron-density $N$, wave-frequency, direction of earth's magnetic field, the direction of incidence upon the ionosphere. The four roots of the quartic for $q$ correspond to the upgoing-downcoming ordinary/extraordinary magneto-ionic components.

"It is shown that, to plot the four roots of $q$ as a function of $N$ for fixed values of the other parameters, it is easier to restate the fundamental formula as a cubic equation for $N$ as a function of $q$. Using this cubic equation, simple limiting curves between which the actual curves necessarily lie are deduced, thereby facilitating plotting in a particular case. Methods for mass-production of such propagation-curves are devised but not applied on a large scale.

"Formulas for attenuation, lateral deviation, horizontal range, and equivalent paths of magneto-ionically split wave-packets are derived and illustrated graphically. The amount to which a wave-packet can be permanently deviated out of its plane of incidence by the earth's magnetic field can be extremely large, but probably not in cases of practical importance in commercial radio communication."

621.396.11 715

**The Reflection Coefficient of a Linearly Graded Layer.**—G. Millington. (*Marconi Rev.*, Oct.–Dec. 1949, Vol. 12, No. 95, pp. 140–151.) A study of the reflection of a horizontally polarized plane wave incident obliquely on a layer in which there is a linear variation of dielectric constant from a constant value on one side of the layer to another constant value on the other side of the layer. The curves given have particular application to the propagation of u.h.f. radio waves in the troposphere, and should be a useful contribution to the understanding of u.h.f. long-distance transmission.

The curves refer only to horizontal polarization, but when the change in refractive index is very small and the wave is at nearly grazing incidence, the reflection coefficients for vertical and for horizontal polarization are approximately equal numerically, though their analytical forms are different. Under these strictly limited conditions, the curves may also give the numerical value of the reflection coefficient for vertical polarization.

621.396.11 : 523.72 716

**Observations of the Propagation of Broadcasting Waves, $\lambda = 1\,250$ m, during the Occurrence of Mögel-Dellinger Effects.**—E. A. Lauter. (*Z. Met.*, July 1949, Vol. 3, No. 7, pp. 204–206.) D-layer reflection effects, scattering, and absorption effects for 1 250-m waves are discussed in relation to the occurrence of outbursts of ultraviolet emission from the sun and consequent fading of broadcasting signals. The Mögel-Dellinger

effect does not appear to have been observed hitherto for such long waves. Its occurrence may give new insight into the ionization of the middle stratosphere.

621.396.11 : 535.31 : 551.510.535 717

**Geometrical Optics of the Ionosphere.**—K. Rawer. (*Rev. sci., Paris*, Sept./Oct. 1948, Vol. 86, No. 3298, pp. 585–600.) The dependence of *field strength* upon path length is considered, and a solution is obtained involving the diameter of the ray in the vertical plane containing its path. Calculations are made for the cases of flat and curved earth, and thin and thick ionospheric layers. Diameter/distance curves are plotted. Results obtained for single and multiple reflections show three focal points, (*a*) at the edge of the silent zone, (*b*) for the horizontal ray, (*c*) for the antipodal point. Approximations for the case of a parabolic layer are considered and discrepancies between these and the results of Försterling & Lassen (1932 Abstracts, pp. 87 & 217) and of Appleton & Beynon (3290 of 1940) are noted. The application of the curves finally obtained to propagation problems is briefly considered.

621.396.11 : 551.510.535 718

**Ray Paths of Radio Waves in the Ionosphere.**—H. Poeverlein. (*Z. angew. Phys.*, Oct. 1949, Vol. 1, No. 11, pp. 517–525.) Detailed discussion of the theoretical bases for the graphical construction of ray paths. Several *graphical* representations of the dependence of the refractive index on the direction of the wave normal are given and discussed.

621.396.11 : 551.510.535 719

**Some Problems of Wave-Diffraction in the Ionosphere.**—J. A. Ratcliffe. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, p. 288.) Summary of Pennsylvania State College Conference paper. "Simultaneous observations of radio 'fading' at two points force us to the conclusion that radio waves are returned from the ionosphere by a process of diffractive reflection as if from a rough surface. The diffraction effects produced by a simple model of this kind are considered and are related to well-known phenomena of optical diffraction. It is shown that the methods of calculation are similar to those employed in the analysis of random fluctuations such as occur with radio 'noise'. The fading is shown to be produced by movements of the random irregularities in the ionosphere and is related to their turbulent or regular motions. It is shown how winds in the ionosphere can be measured. Analogies are made with some well known optical phenomena."

621.396.11 : 551.510.535 720

**Oblique Incidence Propagation Work of the Central Radio Propagation Laboratory.**—R. Silberstein. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, p. 288.) Summary of Pennsylvania State College Conference paper. Discussion of experiments, begun in 1947, using a 13.7-Mc/s transmitter capable of an output of several hundred kilowatts. Three methods were used to determine the source of back-scatter; the results indicated that ground-scatter usually predominated over scatter from the E layer. An examination of A-scope photographs confirmed the existence of many of the predicted modes of propagation.

621.396.11 : 551.510.535 721

**The Interpretation of Long Scatter Echo Patterns.**—A. M. Peterson. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, p. 284.) Summary of Pennsylvania State College Conference paper. The mechanism of ground scatter is analysed. It is shown that angle focusing and time-delay focusing produce characteristic patterns on pulse-echo equipment. It is deduced that E-region ionization contributes relatively little to long scatter.

621.396.11 : 551.510.535 722

**Pulse Transmission over Long Distances.**—J. A. Pierce. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 282–283.) Summary of Pennsylvania State College Conference paper. The two methods used for recording fixed-frequency pulse signals received from a distant transmitter are discussed. These are (*a*) the use of stable oscillators to control the recurrence period; (*b*) the use of a servomechanism to synchronize the controlling oscillator with the received signal. Examples are shown of the records obtained with the synchronizer at frequencies of 9.1 and 12.9 Mc/s, chosen to exhibit the effects of the Pedersen ray. The effects of seasonal changes between July 1948 and June 1949 on the propagation of a 17-Mc/s signal over a 6 000-km transatlantic path are studied, and comparisons made with C.R.P.L. predictions.

621.395.11 : 551.510.535 723

**Effect of the D Ionospheric Layer on Very Low Frequency Radio Waves.**—W. Pfister. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, p. 286.) Summary of Pennsylvania State College Conference paper. The D layer considered is a normal Chapman layer at a height of about 60 km. In long-distance propagation, maximum absorption is observed at 45 kc/s, and the calculated ion concentration is of the order of 100/cm³sec. Reflection coefficients, transmission values and virtual heights are also calculated. Different conceivable forms of ion distribution are considered.

621.396.11 : 551.510.535 724

**The Propagation of Long and Very-Long Waves.**—J. A. Ratcliffe. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 281–282.) Summary of Pennsylvania State College Conference paper. An account of the measurements which have been made in England since the war, to investigate reflection from the ionosphere at steep incidence. Results are described of investigations at frequencies between 16 kc/s, for which the daytime downcoming wave is strong, and 100 kc/s, for which the reflected wave is weak.

The variations in the phase of the reflected wave are recorded for transmissions over different distances; the results provide information about changes in the equivalent height of reflection. An account is given of deductions made from the normal diurnal and annual variations.

A sudden phase anomaly often accompanies a sudden ionospheric disturbance and is observable throughout the disturbance because the amplitude of the very long wave is not seriously diminished by the disturbance. The importance of these anomalies for solar and ionospheric theories is discussed.

Attempts are being made to relate propagation at oblique incidence to propagation at more vertical incidence, both by plotting the ground interference pattern in an aircraft and by making detailed measurements at a point distant 500 km from the transmitter.

621.396.11 : 551.510.535 725

**Sporadic Ionization at High Latitudes.**—J. H. Meek. (*J. geophys. Res.*, Sept. 1949, Vol. 54, No. 3, pp. 284–285.) Summary of Pennsylvania State College Conference paper. Discussion of the variations in the $(h', f)$ trace produced by different types of sporadic ionization recorded in northern Canada. Estimates of the speed of motion of sporadic ionized clouds range from 400 km/hr to 1 200 km/hr for virtual heights from 100 km to 300 km.

621.396.11 : 551.510.535 726

**Summary of Symposia** (at the Pennsylvania State College Conference).—Waynick. (*See* 637.)

621.396.11 : 621.317.324† 727

**The Fresnel Reflection Formula in the Region of Long Waves.**—Roeschen. (*See* 679.)

621.396.11 : 551.594.5 728

**Sky-Wave Observations ($\lambda = 1\,250$ m) during Displays of the Northern Lights in 1947.**—E. Lauter & K. Sprenger. (*Z. Met.*, July 1949, Vol. 3, No. 7, pp. 193–198.) Numerous northern-light observations at German stations on 40 nights during 1947 are correlated with observations at Warnemunde of 1 250-m waves reflected from the E layer. The relation of sky-wave absorption to sunspot relative numbers and to geomagnetic quantities is considered both for evenings with northern-light displays and for those with abnormal E-layer effects.

621.396.11.029.6 729

**V.H.F. and U.H.F. Propagation within the Optical Range.**—M. W. Gough. (*Marconi Rev.*, Oct./Dec. 1949, Vol. 12, No. 95, pp. 121–139.) The mechanism of v.h.f. and u.h.f. propagation can be explained in terms of five major factors, namely (*a*) atmospheric refraction, (*b*) ionospheric reflection, (*c*) tropospheric reflection, (*d*) diffraction, (*e*) ground reflection. (*e*) is the dominant factor in propagation within the optical range, and it is with this aspect of the problem that this article is concerned.

From the ray concept of propagation, a technique has been evolved using simple microwave field strength measurements to forecast the behaviour of v.h.f. and u.h.f. transmissions over a specific optical path. In particular the method predicts, without further measurement, the best height for the receiving aerial on any frequency in the v.h.f. or u.h.f. band.

The process is based on the assumption that ground reflection is confined to a point. This assumption becomes increasingly invalid with increasing wavelength and may lead to errors. However, these can be corrected to a large extent by invoking Fresnel's zone theory, which defines what region of the path is involved in reflection. Furthermore, by examining the ground irregularities in this region the reflecting power of the ground can be assessed.

This article develops the theory of the microwave survey technique and describes and interprets a survey made over a test path, from which the behaviour of specific longer waves is predicted. It concludes with a description of confirmatory experiments made over the same path which verify the predictions.

The 'calibration' of optical paths by the use of microwaves eliminates the necessity for tests on operational wavelengths, and coupled with the lightness and compactness of microwave equipment, saves experimental labour, particularly if the path is to carry transmissions on more than one wavelength.

621.396.81 : 523.78 730

**Field-Strength Observations made during the Total Eclipse of the Sun.**—J. Gross. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, p. 1447.) Curves are given showing the estimated signal strength and readability of WWV s.w. signals received at Eastleigh aerodrome, Nairobi, on 1st Nov. 1948. The transmission path was approximately the same as that of the eclipse. Signals disappeared completely during the totality period and returned to normal by the time the sun was completely unobscured. Signals were of normal strength both on the day before and the day after the eclipse.

621.396.81 : 621.396.67.029.64 731

**The Effect of Antenna Size and Height above Ground on Pointing for Maximum Signal.**—A. H. LaGrone & A. W. Straiton. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1438–1442.) An account of investigations of 3.2-cm transmissions over a 27-mile desert path in Arizona. The effect of aerial size and height above ground on the angle of arrival, as indicated by pointing the aerial for maximum signal, is shown in tables and curves for three measured fields. A comparison is also made of the response of the aerials for various angles of tilt in these fields, and their response in an assumed field made up of two plane-wave components.

621.396.812 732

**Ground-Wave Field-Strength Calculation.**—H. L. Kirke. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, p. 1446.) Comment by G. Millington on 2304 of 1949 and author's reply.

## RECEPTION

621.396.621 733

**Reflex Receiver with Feedback and A.V.C.**—E. G. Beard. (*Philips tech. Commun., Aust.*, 1949, No. 4, pp. 9–14.)

621.396.621 : 621.396.619.13 734

**Sky-Wave F.M. Receiver.**—L. B. Arguimbau & J. Granlund. (*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 101–103.) Early experiments showed that reception in the presence of multi-path signals was much inferior for f.m. than for a.m. When signal and interference (multi-path or other) are of nearly equal amplitude, the limiter must have a uniform response over a very wide band. Thus when two signals differ in amplitude by 5% and in frequency by 150 kc/s, the resultant will sweep through 6 Mc/s, and the limiter must pass all these frequencies equally so as not to introduce a.m. A brief description of circuits and laboratory tests of a receiver is given and it is suggested that a transatlantic f.m. high-fidelity link should be possible if a channel of width 150 kc/s could be spared in the s.w. band.

621.396.623 735

**A De Luxe Receiver with Very High Musical Fidelity: Le Jubilé.**—J. Rousseau. (*T.S.F. pour Tous*, Dec. 1949, Vol. 25, No. 254, pp. 398–404.) An illustrated description of the circuit features of a 33-valve broadcast receiver built by M. Laugier and incorporating various design recommendations which have appeared in the journal in recent years. References to these articles are made on a complete circuit diagram. The receiver has six wavebands and uses three separate amplifier stages for high, medium, and low audio frequencies, and an expansion circuit for contrasts.

621.396.822 736

**Note on Transit-Time Deterioration.**—A. van der Ziel. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, p. 1447.) A simple derivation of some formulae recently published by MacDonald (2408 of 1949). See also 1890 of 1948 (Strutt & van der Ziel).

621.396.822 : 621.317.3 737

**The Application of I.F. Noise Sources to the Measurement of Over-All Noise Figure and Conversion Loss in Microwave Receivers.**—Moxon. (*See* 677.)

621.396.621 738

**Les Récepteurs de Radiodiffusion (Broadcasting Receivers).** [Book Review]—Y. Angel. Publishers: Eyrolles, 244 pp. (*Radio tech. Dig., Edn franç.*, Dec. 1949, Vol. 3, No. 6, p. 375.)

## STATIONS AND COMMUNICATION SYSTEMS

621.395 + 621.397] : 621.315.212 739

**Progress in Coaxial Telephone and Television Systems.**—L. G. Abraham. (*Bell Syst. tech. Publ. Monogr.* B-1630, 8 pp.; *Trans. Amer. Inst. elect. Engrs*, 1948, Vol. 67, pp. 1520–1527.) A general description of the Bell Telephone L1 system, which provides either 600 telephone circuits or a 2-way television circuit with a bandwidth of 2.8 Mc/s, with details of the coaxial cables used and their associated equipment.

621.395.635 : 621.395.44 : 621.396.931 740

**Vibrating Reed Selective Signaling System for Mobile Telephone Use.**—H. M. Pruden & D. F. Hoth. (*Bell Syst. tech. Publ. Monogr.*, B-1663, 5 pp.; *Trans. Amer. Inst. elect. Engrs*, 1949, Vol. 68, pp. 387–391. Summary in *Elect. Engng, N.Y.*, Nov. 1949, Vol. 68, No. 11, p. 927.) A means of calling a particular car in a group tuned to a common carrier frequency. The transmitter at the fixed station can transmit any combination of four a.f. tones from the 32 available, 10 000 combinations being possible. Each car is equipped with a set of four vibrating-reed selectors (756 below) tuned to selected tones, the combination being different for each car. The operation and advantages of the system are discussed.

621.396.1 : 621.396.931 741

**Radio Communications Services: Part 2.**—(*FM-TV*, July 1949, Vol. 9, No. 7, pp. 18–21, 45.) Continuation of 2907 of 1949.

621.396.1 : 621.396.931 742

**New Frequency Assignments for Mobile-Radio Systems.**—G. H. Underhill. (*Elect. Engng, N.Y.*, Nov. 1949, Vol. 68, No. 11, pp. 951–955.) Description of the method by which a new operating frequency will be selected for each of 500 mobile-radio systems.

621.396.619.13 743

**On the Question of the Applicability of Single-Sideband Methods for Frequency-Modulation Transmissions.**—A. Lutsch. (*Fernmeldetech. Z.*, Nov. 1949, Vol. 2, No. 11, pp. 347–351.) A theoretical treatment of the subject shows that no simple demodulation process can be applied. Investigation of the a.m. which occurs in s.s.b. systems indicates that the phase swing is limited to $\pm 1.7$ radians, so that only narrow-band modulation is possible. Formulae are given for determining the behaviour of the demodulated l.f. curves, and distortion is discussed. From these considerations, s.s.b. f.m. appears to be impracticable for communication purposes.

621.396.619.13 : 621.396.712 : 534.86 744

**A Demonstration of Experimental Broadcast Transmitting and Receiving Equipment.**—Honnor. (*See* 535.)

621.396.65.029.63 745

**A Wide-Band V.H.F. Radio-Relay System.**—W. S. McGuire. (*Proc. Instn Radio Engrs, Aust.*, June 1949, Vol. 10, No. 6, pp. 160–165.) Paper presented at the Australian I.R.E. convention, Sydney, 1948. The link operates in the frequency range 152–165 Mc/s, the frequency deviation being $\pm 30$ kc/s. Frequency response is constant to within $\pm 1$ db from 50 c/s to 17 kc/s. Terminal and repeater equipment are described, with performance figures for an experimental 22-mile link.

621.396.712+621.396.97 746

**High-Frequency Broadcasting in Australia.**—N. S. Smith. (*Proc. Instn Radio Engrs, Aust.*, Oct. 1948, Vol. 9, No. 10, pp. 4–20.) Paper presented at the Australian I.R.E. Convention, Nov. 1948, and giving a historical outline of the development of broadcasting in the frequency band 3–30 Mc/s. The transmission characteristics of the ionosphere, as they affect this band, are summarized. 'Internal' and 'external' services are dealt with in two main sections and brief technical summaries are included of the transmitters, the power output of which ranges from 1 to 100 kW. Descriptions are given of aerial arrays and switching arrangements.

621.396.72 : 621.3.018.4(083.74) 747

**Experimental Standard-Frequency [3-kW] Transmitting Station, WWVH.**—G. H. Lester. (*Communications*, Sept. 1949, Vol. 29, No. 9, pp. 20–23, 33.) See also 1982 of 1949.

621.396.931 : 621.317.3(083.74) 748

**Standards on Railroad and Vehicular Communications: Methods of Testing, 1949.**—(*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1372–1375.) Prepared by an I.R.E. committee.

621.396.931 : 621.396.619.13 749

**South Australian Police F.M. Network.**—R. W. Goss. (*Communications*, Sept. 1949, Vol. 29, No. 9, pp. 14–16.) General description of the 72.5-Mc/s system, which comprises a 250-W transmitter located in Adelaide, a 25-W remote station 1 250 ft above sea level, and 10-W mobile units. Points within 25–30 miles of the 250-W transmitter are covered.

621.396.931 : 621.396.619.13 750

**Mobile F.M. Communications Equipment for Australian Conditions.**—R. A. Ratcliffe & R. S. Zucker. (*Proc. Instn Radio Engrs, Aust.*, April 1949, Vol. 10, No. 4, pp. 101–113. Correction, June 1949, Vol. 10, No. 6, p. 159.) Paper presented at the Australian I.R.E. convention, Sydney, 1948. Examples of performance specifications for mobile f.m. v.h.f. equipment, issued by various public bodies in Australia, and also for typical American installations are tabulated and discussed. A standard performance specification for Australian equipment is proposed. A description is given, with photographs and circuit diagrams, of mobile transmitter and receiver units designed to meet the requirements of the proposed specification.

621.396.931 : 621.396.619.13 751

**V.H.F. Mobile Communication Systems.**—A. J. Campbell. (*Proc. Instn Radio Engrs, Aust.*, March 1949, Vol. 10, No. 3, pp. 73–82.) Paper presented at the Australian I.R.E. convention, Sydney, 1948. The main factors which have encouraged the development of such systems are reviewed. A detailed description is given of a standardized range of f.m. equipment designed to provide reliable service over an area of at least 20-miles radius around a central station. Carrier frequencies in the bands 70–80 Mc/s and 152–166 Mc/s are used, with crystal control of the transmitter frequency. The double-superheterodyne receivers derive both local-oscillator frequencies from appropriate points in a crystal-driven multivibrator chain, the crystals vibrating in the third harmonic mode and thus providing initial high frequencies which greatly simplify design. Various types of aerial are available. Two selective-calling systems are mentioned: (*a*) a 2-tone system for a maximum of 33 mobile units, in which mechanically resonant reed selectors are fitted; (*b*) a digit system similar in principle to that of the automatic telephone. This has a much greater capacity than the 2-tone system. With a 2-digit code, about 80 mobile units can be called.

621.396.933 752

**The Modern Civil Aircraft Radio Station.**—S. C. Wallace. (*Marconi Rev.*, Oct./Dec. 1949, Vol. 12, No. 95, pp. 152–156.) The methods of installing radio instruments in aircraft are reviewed. The general trend towards standardization has now developed into the 'standard racking scheme'. This is described and its use illustrated.

## SUBSIDIARY APPARATUS

621-526 : 623.451 753

**German Missile Accelerometers.**—T. M. Moore. (*Elect. Engng, N. Y.*, Nov. 1949, Vol. 68, No. 11, pp. 996–999.) Control of steering and fuel cut-off in German rocket missiles was effected by various methods based on the measurement and integration of acceleration. Single integration systems used gyroscope

precession and electro-deposition. A circuit for double integration by electro-deposition methods is described. A system for lateral control, using an electrically-maintained moving-coil vibrator, is also described; by triple integration of lateral acceleration the missile was made to return to its course in spite of side winds.

621.314.5 : 621.526 754
**An Electronic D.C. to A.C. Converter for Use in Servo Systems.**—E. E. St. John. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1474–1478.)

621.314.632 755
**The Easy-Flow Characteristics of Copper Oxide Rectifiers.**—F. Rose & E. Spenke. (*Z. Phys.*, 30th Aug. 1949, Vol. 126, Nos. 7/9, pp. 632–641.) The present diffusion theory of dry rectifiers gives the d.c. characteristics only qualitatively. An improvement in measurement technique and a more exact interpretation of the theory shows that, for $Cu_2O$ rectifiers, large portions of the easy-flow characteristic are given quantitatively by the theory. The existence of the chemical barrier layer is confirmed. See also 826 of 1942 (Schmidt).

621.325.035 756
**Vibrating Reed Selectors for Mobile Radio Systems.**—A. C. Keller & L. G. Bostwick. (*Bell Syst. tech. Publ. Monogr.*, B-1662, 4 pp.; *Trans. Amer. Inst. elect. Engrs*, 1949, Vol. 68, pp. 383–386.) The vibrating reeds are attached to a metal block to form a small tuning fork. A pole-piece attached to a small magnet is arranged centrally between the tines and an actuating coil surrounds both tines and pole-piece. When the coil is energized with the correct frequency and voltage, the tines vibrate, thus causing a contact to close during a fraction of each cycle of vibration. Another type, without contacts, can be used to provide sharply selective filters. Performance data, graphs, and photographs are included.

621.396.622 757
**Theoretical and Experimental Study of Detection by Silicon Crystals.**—Lapostolle. (*See* 788.)

621.396.68 : 621.397.62 758
**Power Supplies for Home Television Receivers.**—V. Wouk. (*Elect. Engng, N.Y.*, Dec. 1949, Vol. 68, No. 12, pp. 1061–1065.) Discussion, with circuit diagrams, of various methods of supplying (*a*) l.v. B+ and bias power, (*b*) h.v. power for the c.r. tube.

778.3 : 621.317.755 759
**Techniques in High-Speed Cathode-Ray Oscillography.**—C. Berkley & H. P. Mansberg. (*J. Soc. Mot. Pict. Engrs*, Nov. 1949, Vol. 53, No. 5, pp. 549–578.) Photographic technique for recording c.r.o. traces of (*a*) recurrent, (*b*) transient, (*c*) drifting phenomena is discussed.

## TELEVISION AND PHOTOTELEGRAPHY

621.397.335 760
**A Television Synchronising-Signal Generator.**—J. E. Benson, H. J. Oyston & B. R. Johnson. (*Proc. Instn Radio Engrs, Aust.*, May 1949, Vol. 10, No. 5, pp. 128–139.) Paper presented at the Australian I.R.E. convention, Sydney, 1948. The synchronizing signal waveforms of the British and American television systems are described, and the technical requirements of the generating equipment are indicated. An outline is given of some typical circuit techniques used in synchronizing generators, and an experimental 525-line 25-frame synchronizing-signal generator unit is described which uses equalizing pulses and is suitable for negative modulation.

621.397.5 761
**Distant Electric Vision.**—J. D. McGee. (*Proc. Instn Radio Engrs, Aust.*, Aug. 1949, Vol. 10, No. 8, pp. 211–223.) 1948 Australian I.R.E. Convention lecture. A historical review of television, with special reference to the development of the emitron, super-emitron and C.P.S. emitron, and discussion of their relative merits. (*Note.* This lecture was recorded in England on magnetic tape and reproduced, with slides, in Sydney. The tape, slides and original manuscript were sent from England by air mail.)

621.397.5 762
**Dr. Lee de Forest's Color Television System.**—(*Tele-Tech*, Nov. 1949, Vol. 8, No. 11, pp. 41–42.) Tricolour filters, made up of hexagonal elements of the three primary colours are made to oscillate synchronously before the camera and the reproducing c.r. tube, so that every element of the transmitted picture is scanned by each of the primary colours every twentieth of a second.

621.397.5 763
**New Directions in Color Television.**—D. G. F. (*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 66–71.) Comparison of the field-sequential, line-sequential and dot-sequential methods of transmission. Principles of the C.B.S., C.T.I. and R.C.A. systems are discussed.

621.397.5 : 535.88 764
**An Experimental Large-Screen Television Projector.**—P. Mandel. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1462–1467.) Description of equipment for projection on a special screen of the directional type. The screen, of dimensions 3 m × 2.25 m, is an assembly of about 15 000 elementary screens, each consisting of 200 very small spherical mirrors stamped in a highly polished Al sheet. See also 2941 of 1948.

621.397.61 : 621.396.619.23 765
**Mid-Level Modulation for TV Transmitters.**—N. H. Young. (*Communications*, Sept. 1949, Vol. 29, No. 9, pp. 10–11.) Modulation is applied to the grid of the penultimate r.f. amplifier and the vestigial sideband characteristic is provided by a filter between the transmitter and the aerial. The size and cost of the modulator system are thus reduced.

621.397.62 766
**Influence of the Choice of a Television Standard on on the Construction of Receivers.**—M. Chauvierre. (*Radio franç.*, Nov. 1949, No. 11, pp. 7–17.) Discussion with particular reference to receiver cost and quality of reception. Part 1 deals with the effect of synchronization and modulation methods on image stability. In part 2 the choice of the line standard is considered with respect to receiver cost. Part 3 presents the results of comparisons of the definition theoretically possible with that actually observed for about 1 000 receivers, 600 in the U.S.A., 100 in England and 300 in France. The majority of the receivers gave definition much inferior to that corresponding to the quality of the transmission. From the point of view of economical receiver construction, positive video modulation with a.m. for the sound channel is preferable. The line standard should be in the range from about 400 to about 600 lines. For very high definition in de-luxe receivers and for large-screen projection a standard of over 1 000 lines appears indispensable.

621.397.62 767
**Study of a Sound Receiver for Television.**—H. Gilloux. (*Radio prof., Paris*, Nov. 1949, Vol. 18, No. 178, pp. 14–16.) The miniature receiver described, working on 42 Mc/s, gave satisfactory results at distances up to 100 km from a standard transmitter. It uses rimlock

valves and has a normal sensitivity of 200 μV, using an EF42 h.f. stage with an input impedance of 75 Ω. Circuit details and constructional features are noted.

621.397.62 768
**The Video-Frequency Stage.**—R. Gondry. (*Télévis. franç.*, Nov. 1949, No. 53, pp. 6–9.)

621.397.62 : 621.318.4 769
**Coils for Television Receivers.**—F. Juster. (*Télévis. franç.*, Nov. 1949, No. 35, pp. 15–18, 24.) Graphical methods for designing h.f., i.f. and video-frequency coils with maximum *Q*.

621.397.62 : 621.396.615.17 : 621.385.2 770
**The Efficiency Diode in Television Line Time Bases.**—Coxall. (*See* 782.)

621.397.82 : 621.326 771
**TV Interference from Incandescent Lamps.**—D.G.F. (*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 132–148.) Intermittent contact of a lamp filament with its supports may produce short r.f. pulses in the 50–100-Mc/s region, which repeat at power-line frequency.

621.397.5 772
**Basic Television.** [Book Review]—B. Grob. Publishers: McGraw-Hill, New York, 1949, 596 pp., $5.00. (*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 232, 234.) The author "has managed to present the entire field of television in a completely understandable and logical way which should appeal to the technician who is sincerely interested in what really goes on behind the picture tube".

## TRANSMISSION

621.316.726 773
**Simplified Frequency Stabilization.**—O. A. Tyson. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, p. 1445.) About 1% of the output of a klystron is diverted by a directional coupler to a frequency discriminator comprising an unmatched hybrid junction, a phase changer, and a resonant cavity. The phase changer is adjusted to deliver equal modulated power to the detectors terminating the E and H arms of the hybrid when the cavity is detuned and the a.f.c. switch is 'off'. The cavity is then tuned to resonance, which again equalizes the power in the E and H arms, and the a.f.c. switch is turned to 'on'. Any subsequent frequency drift unbalances the system and the amplified and rectified detector outputs can be used to control the modulator valve and hence the amplitude of the square-wave output, which is applied to the steady-state klystron reflector voltage by derivative coupling, with the result that the oscillation frequency is stabilized. The scheme, with slight modification, can be made to work either with c.w. or a.m. output. Stabilization to within about 1 part in $10^3$ is obtained.

621.396.61 774
**Citizens' Band Transmitter.**—A. R. Koch. (*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 118, 132.) Describes the construction of a small transmitter giving an output of 7 W at 450 Mc/s. The tripler and power-amplifier stages use capacitor-tuned parallel-plate resonant lines with lighthouse valves. For other articles on Citizens' Radio see 3482 of 1949 (Lurie) and back references.

621.396.61 : 621.396.97 775
**Co-Ordinated Design of A.M. Broadcast Transmitters for a Range of Power Output.**—P. R. Hellyar. (*Proc. Instn. Radio Engrs, Aust.*, July 1949, Vol. 10, No. 7, pp. 181–189.) 1948 Australian I.R.E. convention paper. A basic design of m.f. (540–1 600 kc/s) a.m. broadcast transmitters, for the output range 200 W–60 kW, is considered with particular reference to Australian requirements. Standardization of components and methods gives advantages in both design and manufacture without sacrifice of performance or appearance. Details of several standardized transmitters are included.

621.396.61.029.56/.58 776
**S.F.R. 10-kW Short Wave Transmitter.**—H. Grumel. (*Ann. Radioélect.*, Oct. 1949, Vol. 4, No. 18, pp. 344–357.) An illustrated description of a standardized communications transmitter designed for telegraphy and telephony operation in any climate on heavy-traffic commercial channels. Separate h.f. units provide for simultaneous operation in three bands which together cover the frequency range 2.5–23 Mc/s. Frequencies are crystal controlled and power input, from 230–400-V 50-c/s supply is 30, 75 or 90 kVA according to the equipment combinations in use.

621.396.61.029.56/.58 777
**S.F.R. 2-kW and 20-kW Single-Sideband Transmitters.**—G. Pembose. (*Ann. Radioélect.*, Oct. 1949, Vol. 4, No. 18, pp. 358–371.) An illustrated description, with block and circuit diagrams. The advantages of s.s.b. operation are outlined and the application of the normal suppression method to these transmitters is discussed. The use of quartz filters reduces the number of frequency conversions necessary to three, which are effected at 84 kc/s, at 2 520 kc/s and at a variable frequency. Each transmitter has a frequency range of 3.75–23 Mc/s and includes an automatic quick-action frequency-selection device. C.w. and a.m. telegraphy, and telephony operation can also be arranged; in s.s.b. working a 'pilot frequency' signal is transmitted for demodulation purposes at the receiver.

## VALVES AND THERMIONICS

621.383.4 778
**A New Germanium Photo-Resistance Cell.**—J. N. Shive. (*Phys. Rev.*, 15th Aug. 1949, Vol. 76, No. 4, p. 575.) A description of a device using a photoconductive property of germanium which combines high spatial resolving power with an over-all quantum efficiency greater than unity. The electrode geometry is similar to that of the double-surface transistor. The characteristics of the collector are modified by the photoliberation of charge in the neighbouring semiconductor. The responsive area on the germanium is confined to a small region on the illuminated side immediately opposite the collector contact. Specimen static characteristics are given showing load power responses of the order of several tenths of 1 mW per millilumen. The cell response to different light modulation frequencies is flat up to the highest frequency studied, about 200 kc/s. The response rises slowly from the visible yellow to a peak at 1.5 μ and decreases rapidly beyond 1.6 μ.

621.385.032.213 : 537.581 779
**Thermionic Electron Emission from Carbon.**—Ivey. (*See* 604.)

621.385.2 780
**Cathode Field in Diodes under Partial Space-Charge Conditions.**—H. F. Ivey. (*Phys. Rev.*, 15th Aug. 1949, Vol. 76, No. 4, pp. 554–558.) A method is given for calculating the electric field at the cathode of a plane or cylindrical diode as a function of diode current. The results are believed to be applicable to any electrode geometry and to the study of thermionic cathodes in the Schottky emission region. The effect of anode voltage on cathode field, and the potential distribution in a plane diode are also discussed.

621.385.2 : 537.525.92 781
**Diode Space Charge for Any Initial Velocity and Current.**—L. Page & N. I. Adams, Jr. (*Phys. Rev.*, 1st Aug. 1949, Vol. 76, No. 3, pp. 381–388.) "The space charge equation is solved for any initial velocity which is the same for all electrons and for any realizable

current, for (*a*) the plane diode, (*b*) the cylindrical diode. For the plane diode it is shown that all cases can be obtained from a single pair of master curves. For the cylindrical diode a series valid for large distances from the axis is obtained first, the form of which shows that the complete group of solutions form a bounded set. Next three solutions valid for small distances from the axis are obtained by means of which the first set of solutions can be extended. In addition to master curves representing these solutions, curves are plotted for special values of the initial velocity and current."

621.385.2 : [621.397.62 : 621.396.615.17] 782

**The Efficiency Diode in Television Line Time Bases.**—N. Coxall. (*Philips tech. Commun., Aust.*, 1949, No. 4, pp. 3–8.) Description of the miniwatt diode Type EA40, with emphasis on the increase in line timebase efficiency and on the circuit simplification possible when this diode is used.

621.385.3 783

**New Triodes for Metre and Decimetre Waves.**—L. Liot. (*Radio franç.*, Nov. 1949, No. 11, pp. 3–6.) A description is given of the Type T 30H disk-seal triode constructed by the S.A.D.I.R. The electrodes are cylindrical and the valve is used in a vertical position. Maximum anode and grid voltages are + 1 000 V and − 200 V respectively. Maximum anode dissipation is 30 W and the upper frequency limit about 700 Mc/s. The construction details are given of a symmetrical twin-line oscillator using two such valves. Circuit characteristics and operating conditions are given for four frequencies in the range 220–490 Mc/s.

621.385.38 784

**Pulse-Controlled Thyratron.**—J. G. Skalnik. (*Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 120–168.) A short positive pulse of variable phase with respect to the 60-c/s anode voltage of a positive-grid thyratron controls its conduction period.

621.385.83 : 538.691 785

**Study of the Magnetic Focusing of Cylindrical Electron Beams.**—G. Convert. (*Ann. Radioélect.*, Oct. 1949, Vol. 4, No. 18, pp. 279–288.) See 485 of February.

621.385.83 : 621.396.615.142 : 538.691 786

**Magnetic Focusing of a Cylindrical** [electron] **Beam with Density Modulation.**—R. Berterottière. (*Ann. Radioélect.*, Oct. 1949, Vol. 4, No. 18, pp. 289–294.) The transverse dispersive forces in an irregular electron beam are considered, and the behaviour of such a beam in a compensating magnetic field is discussed. Space-charge effects lead to considerable values of the radial debunching force, so that a magnetic field variable along the beam according to a certain law is necessary. Oscillations of electrons situated between two packets are minimized by using a sufficiently strong field.

621.396.615.142.2 787

**Beam-Loading Effects in Small Reflex Klystrons.**—W. W. Harman & J. H. Tillotson. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, pp. 1419–1423.) Beam-loading $Q$ is determined from separate measurements of $Q$ for valve and cavity (*a*) without and (*b*) with the accelerating voltage applied. Loading effects are much greater than predicted by published analyses; they vary approximately linearly with total oscillator load, a variation not previously noted. Analysis of loading effects produced by secondary electrons ejected into the intergrid space by the main electron stream shows the beam-loading to be large, with linear variation with load, and indicates the possibility of negative beam-loading.

621.396.622 788

**Theoretical and Experimental Study of Detection by Silicon Crystals.**—P. Lapostolle. (*Onde élect.*, Dec. 1949, Vol. 29, No. 273, pp. 429–448.) An account of research, extending over a period of two years, on a selection of detectors of British and German origin, together with full discussion of a theory based on the classical equivalent circuit. The method of showing the voltage/current characteristic on the screen of a c.r.o. is described and static characteristics for actual detectors are given. Methods of determining the impedance of detectors at low frequencies and at frequencies up to 10 Mc/s are described and rectification characteristics in the range 30–100 Mc/s and at 500 Mc/s are shown. Impedance and power measurements at decimetre wavelengths are considered in detail. Experimental results for powers above 10 mW are compared with values calculated from the theory given. Agreement is satisfactory and observed variations can be interpreted correctly by the theory, which indicates a slow decrease of detector sensitivity with increasing frequency and stresses the importance, even at v.h.f., of a very high backward resistance.

621.396.645 : 537.311.33 : 621.315.59 789

**The Transistron Triode Type P.T.T.601.**—R. Sueur. (*Onde élect.*, Nov. 1949, Vol. 29, No. 272, pp. 389–397.) Discussion of the properties of Ge, the effects at the point of contact between a conductor and a semiconductor, and the mechanism, construction and applications of the transistron. See also 2078 of 1949 (Aisberg).

621.396.645.029.4/.52 : 621.396.822 790

**Selective Amplification at Low Frequencies.**—de Queiroz Orsini. (*See* 596.)

621.396.822 791

**Spontaneous Fluctuations in Double-Cathode Valves.**—K. S. Knol & G. Diemer. (*Wireless Engr*, Oct. 1949, Vol. 26, No. 313, p. 345.) Measurements at 6 Mc/s on a valve with two indirectly heated plane cathodes 1 mm apart, and on a second valve with directly heated tungsten filaments of diameter 100 $\mu$ and 0.5 cm apart, do not support Fürth's theory (2418 and 2419 of 1948).

621.396.822 792

**Note on Transit-Time Deterioration.**—A. van der Ziel. (*Proc. Inst. Radio Engrs, W. & E.*, Dec. 1949, Vol. 37, No. 12, p. 1447.) A simple derivation of some formulae recently published by MacDonald (2408 of 1949). See also 1890 of 1948 (Strutt & van der Ziel).

537.533 793

**L'Émission Électronique.** [Book Review]—J. Bonchard. Publishers: Librairie de la Radio, 160 pp. (*Radio tech. Dig., Édn franç.*, Dec. 1949, Vol. 3, No. 6, p. 379.) "A book which should interest radio technicians, particularly those concerned with the manufacture of transmitting and receiving valves, as well as students and physicists."

## MISCELLANEOUS

061.3 : 621.3 794

**A.I.E.E. Fall General Meeting Conference Papers Digested.**—(*Elect. Engng, N.Y.*, Dec. 1949, Vol. 68, No. 12, pp. 1091–1098.) Authors' summaries of 27 papers read at the meeting.

621.396 795

**I.R.E.** (Australia) **Radio Engineering Convention.**—(*Proc. Instn Radio Engrs, Aust.*, Oct. 1948, Vol. 9, No. 10, pp. 21–27.) Summaries of the technical papers presented at the convention held in Sydney, Nov. 1948.

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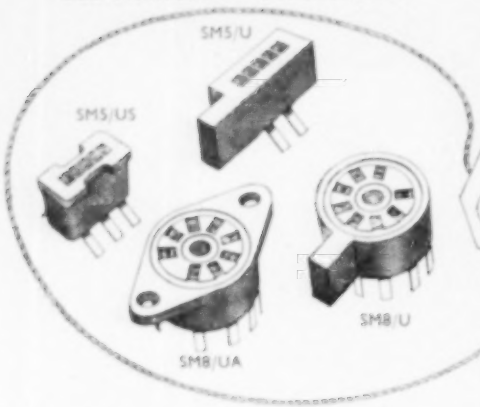


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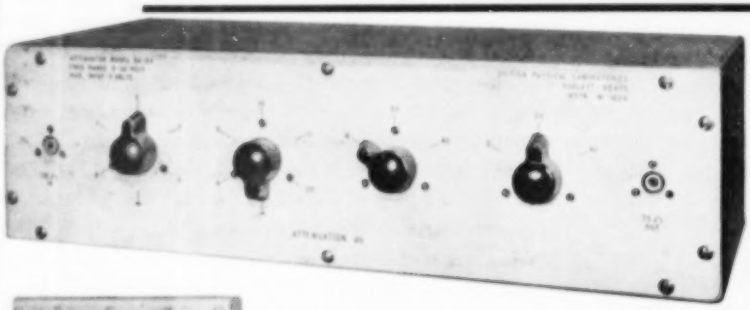

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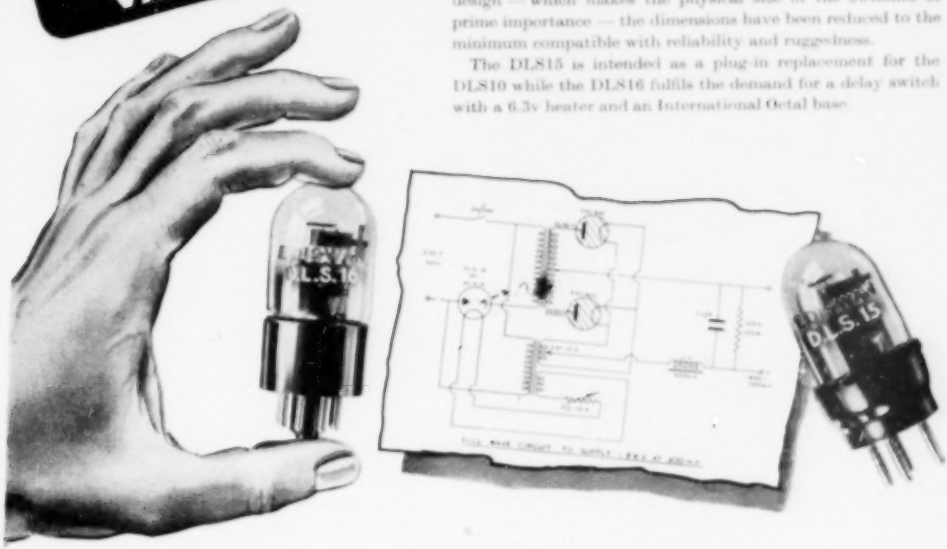


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| Filament Current (amps) ... ... ... | 0.75* |
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*(Delay time may be increased to 180 secs. by variation of filament voltage)*

Maximum Peak Current (low voltage rating) 5 amps at 240v.

Maximum Peak Current (high voltage rating) 100mA at 1 kv.

Maximum overall length (not including pins) 60 mm.

**at approximately 4 volts*

**DLS16 Rating.**

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|---|---|
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| Filament Current (amps) ... ... ... | 0.48* |
| Delay time at 6.3 volts (secs.) ... | Min. 30 ; Max. 90 |

*(Delay time may be increased to 180 secs. by variation of filament voltage)*

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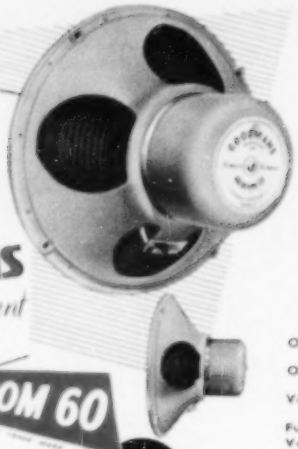

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| Fundamental Resonance | 75c.p.s. |
| Voice Coil Impedance | 15 ohms. |
| Power Rating | 15 watts Peak A.C. |
| Flux Density | 14,000 gauss |
| Total Flux | 158,000 maxwells |
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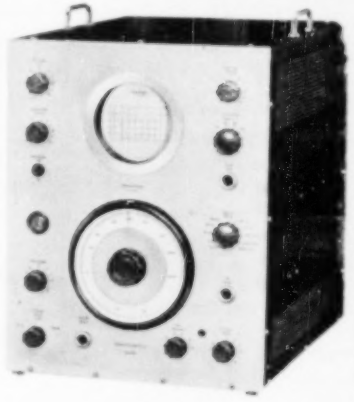

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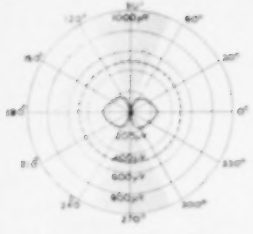

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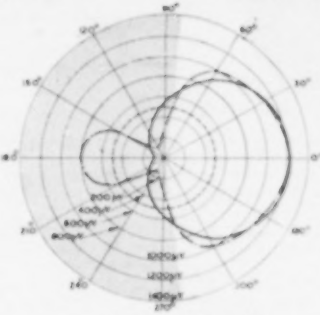
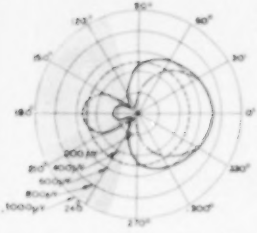
····· DIPOLE DIAGRAM



4-ELEMENT ARRAY COMPRISING λ/4 'H' WITH TWO DIRECTORS SPACED λ/10

—— VISION DIAGRAM
····· SOUND DIAGRAM
····· DIPOLE DIAGRAM



λ/7 'H' LIGHTWEIGHT

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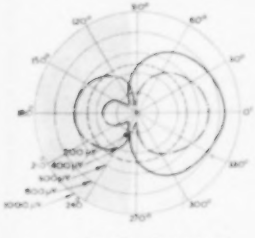
····· SOUND DIAGRAM
····· DIPOLE DIAGRAM



λ/4 'H' SENIOR

—— VISION DIAGRAM
····· SOUND DIAGRAM
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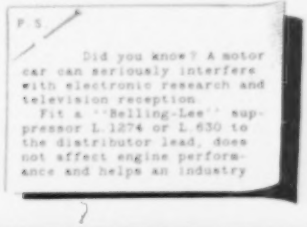
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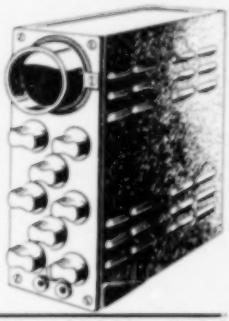

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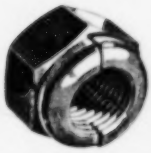

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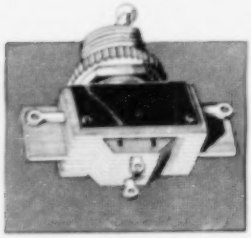

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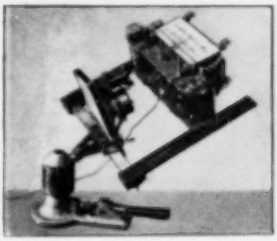

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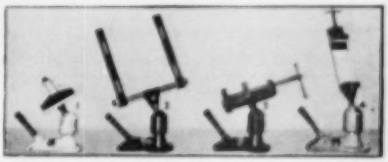

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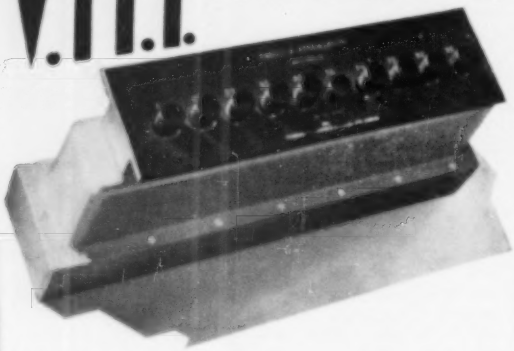

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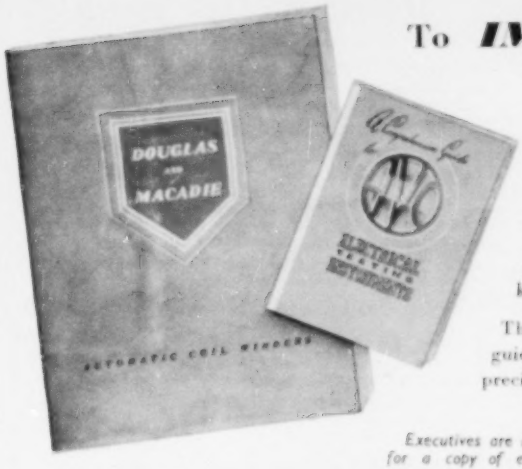


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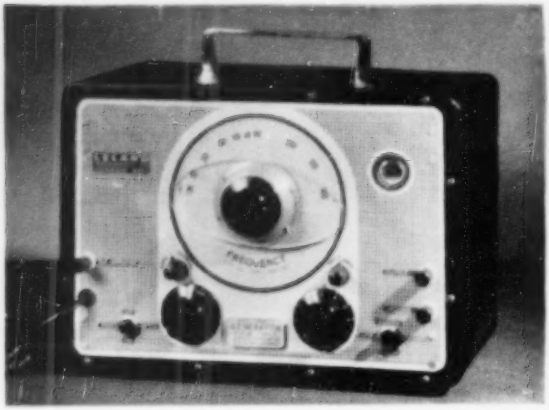

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| 16-16 | 350 | 2½ | 1 | CE34LE |
| 60-100 | 350 | 4⅛ | 1⅛ | CE37LEA |
| 8 | 450 | 1½ | ⅝ | CE18PE |
| 8-8 | 450 | 2½ | 1 | CE34PE |
| 16-16 | 450 | 2½ | 1 | CE35PE |
| 32-32 | 450 | 4⅛ | 1⅛ | CE37PE |

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